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14. ABSTRACT

The charge for this study was to evaluate the feasibility of achieving the goals of the Integrated Topsides program, to identify potential roadblocks in the program and long-term opportunities. We were asked to do this in the context of Navy needs and in the context of contractor proposals for development programs. In the course of our evaluation, we found that many concerns about the multi-function concept, as well as many aspects of proposals to address it were narrowly focused to design concepts intrinsic to traditional approaches. At the same time we were greatly heartened to see the dynamism of research and development in relevant RF technologies, and increasing interest and willingness to adapt commercial developments to military needs.

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Agile Multi-Function Arrays

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Contents

1 EXECUTIVE SUMMARY

The Navy is planning modernization of its ship-borne RF functions, radar, communications, and electronic warfare. The broad concept of the modernization plan is to develop multi-function active phased-array apertures to integrate functions now covered by uncoordinated individual RF systems. The key conclusion of this study is that the plan is not only technically feasible, but is crucial to maintaining technical superiority in a period of rapid commercial developments that have direct application to the Navy's RF functions.

The major impediment to the widespread use of active phased-array apertures has been the costs of the traditional architectures, which in turn have been limited by technical capabilities in digital processing and high-speed electronics. These cost and architecture limitations are changing rapidly with the development of commercial microwave communications and wireless networks. The evolution of these commercial technologies can be exploited in the development of new low-cost phased array systems, if Navy personnel appropriately oversee the development process. Conversely, the wide availability of these commercial technologies poses obvious opportunities for adversaries to develop capabilities that compete or interfere with Navy function. Maintaining Navy technical superiority will require a flexible system, such as that proposed under the Integrated Topside program, to allow rapid adaptation to advances in commercial capabilities.

A synopsis of key technical conclusions, described in more detail in the full study report, follows.

1. The importance of implementing the modernization program as a system cannot be overemphasized. As an example, much of the present impetus for modernization is driven by the ever-increasing electromagnetic interference among shipboard RF functions. Issues of interference can be addressed by combining well-planned design of the physical system with software-controlled scheduling. Planned scheduling should also be used to optimize resource allocation and thus minimize system costs.

2. Commercial competition in developing new device technologies such as phase shifters, power amplifiers and RF-CMOS relevant to the 2 - 18 CHz range under consideration will provide market decisions about cost/performance benefits in the transmit-receive modules (TRM) for phased arrays. Developments of particular interest include commercial CaN power amplifiers and possible integration of III-V channels into Si-CMOS chips. The modernized Navy system should take every possible advantage of the resulting COTS components, allowing custom components only when no alternative can be devised. In the case of specialized components, economies of scale (e.g., use of identical components across all systems and coordination with other military and civilian users of phased array systems) must be pursued. Development of custom capabilities should be pursued only when there is no commercial alternative.

3. The evolution of RF-CMOS for front end electronics will allow increasing digital processing for adaptive beam forming, error correction and predistortion even at the high frequencies needed. Development of these

capabilities in parallel with evolving device technologies will improve performance, help to contain costs, and ultimately will enable new functionality such as spread-spectrum capabilities in electronic warfare.

4. Distributing function among apertures that meet the technical requirements for that function (e.g., power level, frequency band, transmit vs. receive) will be needed to maximize system performance. To the extent possible, individual apertures should continue to support multiple compatible functions. The design cost/function trade-offs to be considered will involve array element allocation and configuration *(e.g., multiple function on many generic elements vs. dedicated elements tailored to function)*. **In** the early stages of system evaluation, design choices for antenna arrays should involve uniform undifferentiated elements that can be rapidly reconfigured, via the standardized interfaces, with respect to functional allocation and associated TRM.

5. The development of the modernized system must include rigorous and stringently enforced standards for open software architecture and vendor-transparent hardware interfaces. This is essential to allow for development in a time of rapidly changing technology. To accomplish this the Navy must actively initiate and oversee consensus-based standards development processes analogous to those used by **IEEE and IETF.**

6. We strongly recommend that the modernization program proceed via spiral development, because making an unchangeable technology choice now would be akin to the historical choice between Beta and VHS video standards. The evolution cycle should proceed sequentially through three stages, where we predict the greatest technical advances will occur during the second of the three stages. The three stages are:

(i) Adapt existing architectures and COTS technology to obtain useful, but possibly limited (e.g., narrow bandwidth or lower power) functionality at the lowest possible costs.

(ii) Develop hybrid systems incorporating technological advances dictated by commercial applications and by progress under stage (i). Exploit increased digital processing to decrease demands on electronic components and increase system functionality.

(iii) Constantly evaluate technology developments for potential to achieve the ultimate functionality of "software-controlled RF," and invest in technology development as needed to pursue the most promising options.

2 INTRODUCTION STUDY CHARGE

2.1 Navy Context

In the past 20 years there has been a rapid proliferation in demand for RF function, e.g., communications, radar and electronic warfare, on naval vessels. As a result, the number of antennas on naval vessels roughly doubled between the 1980s and 1990s. The process has occurred with little central planning, as described in one communication [1]:

"For the most part the present external platform design approach is based on developing separate systems and associated antennas for each individual RF function. The individual antennas are then arranged seeking an optimal solution ... The volume of military messages addressing system blockage and EMI problems, and the expenditures on efforts to mitigate these problems has shown this strategy to be unacceptable."

The problem is so severe that some of the antenna proliferation results from construction of multiple antennas for the same function, in an attempt to circumvent problems with electromagnetic interference (EMI). The end result is an approach to gridlock in carrying out mission RF function, increased radar cross section and increased topside weight. The uncoordinated design of the different systems further causes unnecessarily high training and maintenance costs.

These problems, which in some sense appear to be a mundane housekeeping problem (antenna clutter), in fact stem from a real technology driver, which is the proliferation of capabilities in RF. The Navy's need to address

the house-keeping problem is coupled to a serious need to maintain state-of-the-art-capability and military superiority in RF function. To address these issues, a Innovative Naval Prototype program, Integrated Topsides, is underway to develop a systems approach to shipboard RF function [2]. The keystone of this program is the use of multi-function active phased array (or "electronically-steered array") apertures, with the development of new approaches to mitigate the notoriously high up-front expense of phased array systems [3]. The long-term benefits of such a system are manifest:

- Improve functionality of communications, radar, electronic and information warfare.
- Improve system throughput and adaptability by dynamic scheduling of functions.
- Allow rapid adaptation of waveforms, frequencies, bandwidth to address evolving environments and threats
- Allow constantly evolving technology improvements through the use of an open architecture
- Reduce topside weight, e.g., number of antennas, mechanical drives
- Reduce EMI between systems on the same ship
- Reduce ship radar cross-section
- Reduce installed cost and life cycle cost
- Reduce maintenance and training demands.

Achieving these benefits without the crushing costs that have limited the new destroyer (DDX) program will require creative exploitation of com-

mercial drivers, generating economies of scale by modularity, and optimized system design to maximize functional output per hardware unit.

The focus of this study has been RF functional systems covering the frequency range of 2 GHz to 18 GHz. The ability to address the variety of technical requirements listed is necessary for implementation of different functions on a multi-function aperture. In addition, the mission requirements, such as priorities and time utilization of different functions, must also be addressed in making decisions about which functions can and should be implemented on a shared aperture. For instance, the high-power of Anti-Air Warfare (AAW) radars is a demanding requirement, and thus AAW is likely to be technically inappropriate for implementation on low-cost apertures that are suitable for other low power functions. In terms of functional priority, the urgency of AAW function would limit co-sharing of an AAW aperture to other functions that could be turned off, or rerouted to other apertures at need, without damage to mission requirements.

2.2 Study Overview

The charge for this study was to evaluate the feasibility of achieving the goals of the Integrated Topsides program, to identify potential roadblocks in the program and long-term opportunities. We were asked to do this in the context of Navy needs and in the context of contractor proposals for development programs. In the course of our evaluation, we found that many concerns about the multi-function concept, as well as many aspects of proposals to address multifunction capabilities, were narrowly focused based on design concepts intrinsic to traditional approaches. At the same time we were greatly heartened to see the dynamism of research and development in

relevant RF technologies, and increasing interest and willingness to adapt commercial developments to military needs.

In the following sections we address different aspects of the Integrated Topside program. In Section 3, we begin with a very brief overview of active phased-array concepts. We then address some device development proposals that are narrowly focused to specific problems within traditional transmit/receive module design. The key perspective of Section 3 is that the judging long-term value of such new devices must include evaluation of their market competitiveness. Doing so will help avoid being locked into unsustainable products, and will reduce the costs inherent to use of custom products. In Section 4, we address the evolving benefits of fast semiconductor technology to active phased array systems. This is a time of rapid evolution in capabilities, and we recommend that the Navy allow designs to evolve to exploit developments in the commercial sector. In Section 5, we discuss broader systems aspects. Systems decisions will be strongly influenced by the technology options available, and conversely, the systems requirements will constrain the technologies that can be used. We suggest procedures for maintaining the greatest flexibility in the development program for Integrated Topsides. Finally, in Section 6, we broadly recommend a spiral development program, which allows each subsequent round of technology development to grow based on the lessons of the previous round.

In preparing this report, we were provided with a series of extremely useful briefings from representatives of the Navy, contractors, and academics.

The briefings are listed in Appendix A. The technical issues involved in developing Integrated Topsides draw from many different communities, each with its own set of acronyms. We have attempted to define acronyms where they occur in the text, but also include a list in Appendix B for reference.

3 COST DRIVERS IN TRADITIONAL ACTIVE PHASED ARRAY

3.1 Introduction/Background

Before considering issues associated with combining multiple RF functions into a single system, we begin by reviewing the basic architectures and components in a typical radar system, as shown in Figure 1. A single antenna

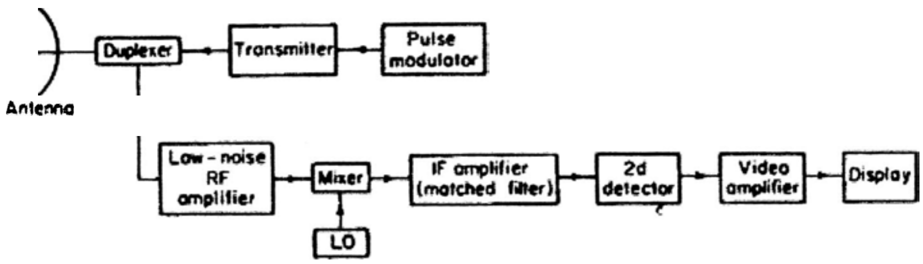


Figure 1: Generic block diagram of components of a radar system. From Reference [5].

or dish is connected to two sets of electronics, the transmit (Tx) and receive (Rx) chains, via a duplexer. (Alternatively, in many cases the transmit and receive dishes, and thus much of the electronics chain, are separate). The function of the duplexer, which can be a circulator or an alternating switch, is to separate outgoing pulses and incoming echoes, directing them into the appropriate Rx/ Tx systems. The transmitting system consists of an RF oscillator, a modulation system which is driven by the control system to produce the pulse or outgoing waveform, and a series of power amplifiers and/or attenuators used to generate the appropriate pulse levels. The receiver usually consists of a super-heterodyne detection system with a low-noise amplifier

(LNA), followed by a mixer which downconverts to a lower frequency (the so-called intermediate frequency, or IF). The down-converted signal is then filtered, further amplified, and finally sent to the detection/display system. Both the detection and pulse shaping are usually digitally controlled, but the entire Tx/Rx electronics chain is otherwise analog. To prevent damage from the high-power outgoing signals and/or interfering signals (RFI), the LNA is often protected at its input by some type of "receiver protector" limiting device (not shown).

In phased array radar, the single radar dish is replaced by an aperture containing an array of small antenna elements. In a passive phased-array radar, by definition having one transmitter per antenna, the general architecture of the front end is very similar to that shown in Figure 1. Between the duplexer and the N elements of the array there is a beamforming (or feed) network, consisting of an N-way power splitter/combiner and individually adjustable phase shifters and attenuators for each radiating element, as shown in Figure 2. Since the elements are driven synchronously, their radiation patterns superpose, resulting in a narrow beam. The relative phasing of the array elements determines the direction in which the superposition is coherent. Because the phases can be electronically controlled[1], this is an electronically-steerable array (ESA), in which the beam can be swept or moved much more rapidly than a large single dish could be rotated. The ability to rapidly reorient the beam allows multiple and/or fast moving targets to be tracked, and is one of the main advantages of phased arrays. The passive phased-array radar of the AEGIS system (the AN/SPY-I) is a good illustration of the improved functionality that a phased array can provide.

[1] An alternative beam-forming approach is to use a Butler network or a Rottman lens, which basically couples N separate input feeds to the N radiators with a fixed phase relationship. Then the N input feeds correspond to N possible beam directions, and beam scanning is performed by mechanically connecting to a different input.

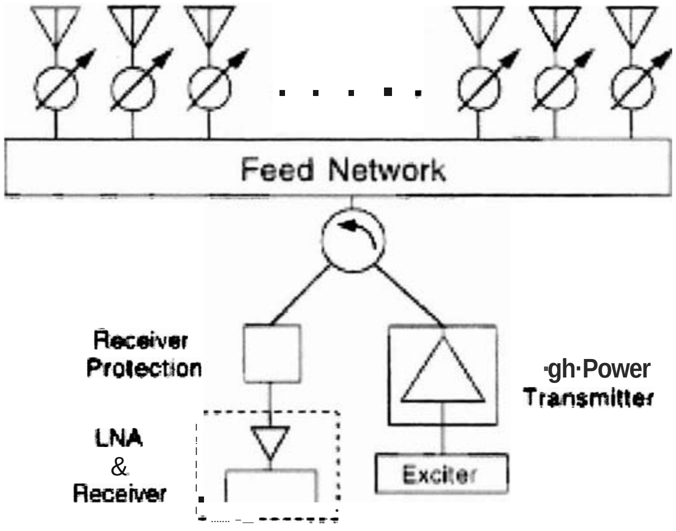


Figure 2: Schematic illustration of the front end of a passive phased array system, showing, from top to bottom, antennas, phase shifters, distribution (feed) network, circulator, and the separate transmit and receive modules. From Reference [6].

In a system with a single dish or radiator, the required directivity or gain increases proportionally to the total effective area of the antenna relative to the wavelength, $G = 4\pi\, A_{\text{eff}}/\lambda^2$, and so determines its size. In an array, the required gain similarly determines the overall size. In order to minimize highly undesirable sidelobes, the array elements must be spaced about half a wavelength ($d = \lambda/2$) apart. This means that the total number of array elements ($N \sim A/d^2 \sim G$) will increase linearly with the array gain. For many radar applications, this gain is 30-40 dB, leading to arrays with typically 1,000 to 10,000 elements. So even though the array will be the same total size, it has many more parts and greater complexity. One advantage of phased arrays is the ability to form and independently steer multiple beams. In a passive phased array, this can only occur via time-sequenced beam transmission. In principle, in an active phased array (active electronically steerable array), as described below, multiple simultaneous beams can

be formed. The key issue in the multi-function phased array program is the extent to which multiple functions can be simultaneously performed in a single aperture to allow reduction of both the number and total area of apertures required on board a ship. The technical concerns include interaction of the different beams due to nonlinearities in amplifiers, dynamic range, electromagnetic interferences, etc.

The array architecture of an Active Electronically-Steered Array (AESA) is shown schematically in Figure 3. In this approach, each radiating element of the array has its own Transmit-Receive Module (TRM), which is essentially a complete analog RF front-end for the radar, connected directly to the elements. Because the output power amplifiers and input LNAs are connected directly to the antennas, this allows the amplifiers to compensate for the inevitable losses in the combiners and phase-shifters of the beam-forming network. In addition, in the active architecture, the phase shifts

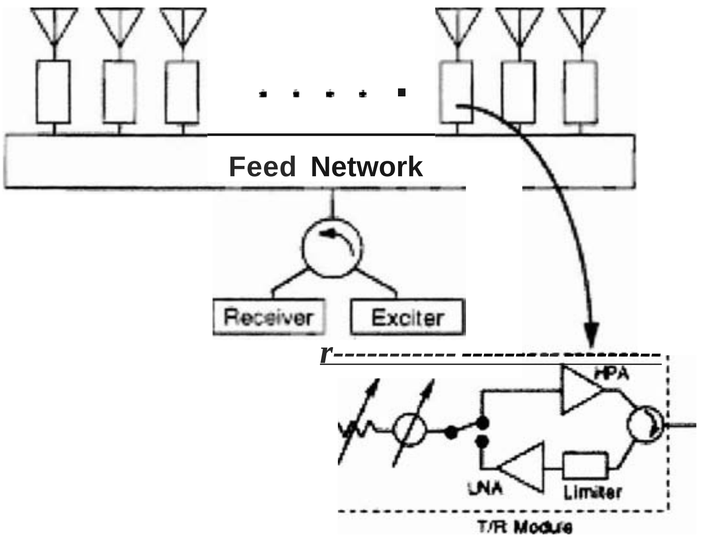


Figure 3: Schematic illustration of the front end of an active phased array system, showing, the integration of the transmit/receive electronics with the individual antenna elements. From Reference [6].

and waveforms applied to each element can be individually tailored, allowing formation of multiple simultaneous beams and sophisticated beam forming operations. The large number and close spacing of sets of RF electronics in the active array means that the TRM modules must be small and power efficient. This has been enabled by use of monolithic microwave integrated circuits (MMICs, usually III-V semiconducting chips) [7]. However, III-V MMICs represent a significant expense when thousands of highly uniform units are required. In addition, a TRM still requires components such as the limiter and duplexer, which use other materials and technologies, and limit the level of integration to a hybrid circuit board or multi-chip module.

The active array architecture has several advantages for combining multiple functions into a single system. Since each TRM in an active array is independently controllable, the array can be dynamically reconfigured or broken into subunits to perform different functions. Even if the entire array is used for multiple signals or beams at once, the power levels in each amplifier are N times smaller than they would be in the electronics of the passive array. This reduces the requirements for saturation power (P1dB) and linearity (IP3) of the individual power amplifiers in the transmit chain illustrated in Figure 4. Similarly, the dynamic range requirements of the LNAs and receiver electronics are reduced since the incoming power per element is reduced. Maintaining the same signal to noise ratio expected in the single-dish or passive array, however, requires the same noise figure (NF) for each LNA in the active array. So while some requirements for the RF electronics are relaxed by adopting the AESA concept, obtaining adequate performance still places stringent requirements on the elements. An alternative approach to a full active array is to break the array into moderately sized subarrays which have their own electronics, and are then further combined afterwards

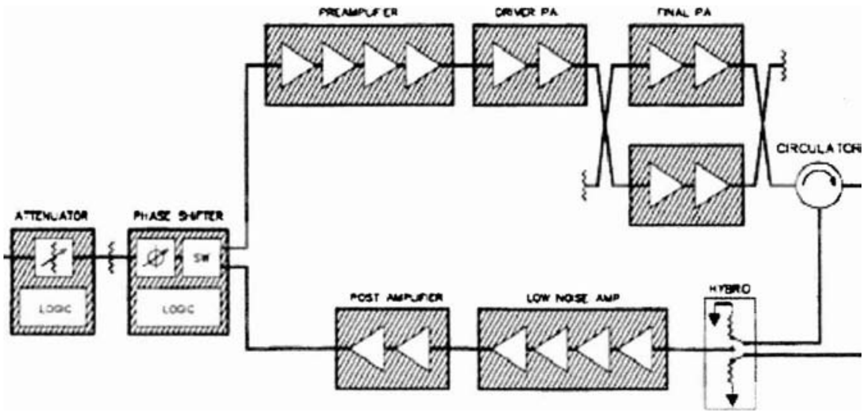


Figure 4: Traditional design of transmit/receive module in an active phased array. The components are typically fabricated in a GaAs MMIC. From Reference [8]. The phase shifter may be either a true time delay, or it may generate a frequency-dependent phase offset. Continuing design improvments include minimizing the parts count, for instance replacing the current cascade of driver and power amplifiers with a single solid state power amplifer. In addition, the addition of integrated tuneable filters can be used to improve dynamic range for wide band signals.

to yield the full array. This is adopted for example in the AN/SPY-1 AEGIS arrays, where sub-arrays of 16 elements each have copies of transmit-receive electronics.

Commercial developments in microwave communication and wireless networks are driving costs down and functional capabilities up for the devices used in transmit/receive modules. As a result, careful selection of COTS components for populating a TRM has significant potential for driving down costs in phased arrays. A major cost component in TRMs has traditionally been the phase shifters. Normally these are fabricated in GaAs as a component of the microwave monolithic integrated circuit [7, 8]. However, there is a possibility that a hybrid package involving phase shifting (either true time delay, or frequency dependent phase offset) using an independent device may yield similar or improved performance at substantially lower cost. This is discussed in more detail in the following section, and digital alternatives are

discussed in Section 4.1. Another issue for Navy applications is the need for extremely high-power in some applications, such as AAW radar, as well as front-end protection of the electronics for the case of high-power jamming or inadvertent illumination via multi-path beams or other unplanned sources. The use of front-end amplifiers based on wide-band-gap GaN is attractive in this respect. Commercial applications, as well as developmental research for military applications, have recently resulted in substantial progress in high-power GaN amplifiers. This is discussed in more detail in Section 3.3. Finally, we discuss in Section 3.4 the issues that arise when wide band signals are desired to allow signals at different frequencies to be handled by the same TRM module. As we will note there, this approach to multifunction must be weighed against the possibility of using parallel electronics designed to handle different frequency bands.

3.2 Beam Forming

Formation or selection of directional RF signals (beam forming) is accomplished by adding variable phase shifts to the signals transmitted or received by all the elements of the array. There are many mechanisms of generating the phase shifts, and here we will generically use the term "phase shifter" to include all mechanisms, including change of the propagation constant along the device path, true-time delay [9] and digital phase shift control. Conventional phased array antennas are designed to create one steered beam at a time, with formation of multiple beams accomplished by time multiplexing the electronics. If one would like to combine the operation of many independent RF channels through a single phased array antenna system, one possibility is to the break the array of elements into subarrays, designing each

to form a different beam as discussed above. Another possibility, that of simultaneously generating signals corresponding to more than one beam at the same antenna element may also be considered using an active phased array system. In each of these cases, hardware implementation requires an independent phase shifter for each simultaneous beam. The possibility of using digitally-generated waveforms to avoid this problem is discussed in Section 4 and Appendix D.

The active phased array antennas needed to support the multiple functions will have to operate over a wide range of frequencies, power levels, as well as responding to demands on angular resolution and duty cycles. In designing a beam forming architecture, one approach would be to create a single wide band front end (TRM) that allows flexible switching of an element between different functions at different times. In this case, the front end design would need to encompass the full bandwidth of all possible signals (the issue of the maximum power level of the most demanding function is discussed in Section 3.3). This would require (in hardware) the use of true time delay phase shifters, that operate by delaying the passage of the RF signal using a variable-length transmission line, or through a related technique. There is a simple and direct connection between their operation and the concept of a phased-array antenna. True time-delay phase shifters operate over a broad range of frequencies, limited at the low end by the length of the transmission line. However, manufacture of MMIC true-time delay phase shifters is a significant cost driver in phased array systems. Alternative device structures, for instance MEMS delay lines warrant further analysis for functionality, cost and reliability. With standardization and volume production, such an alternative could be an economical choice for some implementations of multi-function.

Alternatively, lower cost and greater flexibility may be attained by designing several types of front end electronics, each optimized for a group of functions. As shown in Table ?? above, many of the RF channels proposed for the multi-function phased array system have narrowband signals about a well-defined frequency, for example communications. For this case it may be more appropriate (e.g., improved specific performance at lower cost) to use phase shifters that operate over a relatively narrow bandwidth. Because the RF channels are identified by their frequency, it would be possible run several channels in parallel through a single antenna element, with a different phase shift for each channel, to form independently controlled beams. A straightforward approach to such a system would be to have a separate bandwidth filter and phase shifter for each channel (see for instance Section 5, Figure 23). The question of the ultimate possible cost and performance for a single wide-band front end vs. multiple parallel narrow band front ends remains open. However, at present, narrow band implementations are straightforward and relatively inexpensive. Thus in the spiral development program outlined in Section 6, the use of parallel systems is likely to be important in the first stage of development.

Another interesting possibility is to develop GHz delta-sigma digital-to-analog converters that can directly generate the phase-shifted RF signal over a broad frequency range. This approach is conceptually simple, and it would allow one to generate a wide variety of phase shifting schemes ranging from true time delay to a sophisticated narrow-band phase shifter that controls several channels at once, through a single element. However designing such a system with sufficient resolution and speed will be difficult, and it will require very high digital processing speeds for the digital feed signal (see Section 4.2

and Appendix D). A careful consideration of this approach is needed to see if it is realistic and economical in the near term.

Power level is another consideration. Ideally, the power level of phase shifters should be kept low, to minimize their size and cost. For this reason they are typically placed before the power amplifier, and after the low noise amplifier, as in Figure 3. Alternative placement at the local oscillator of the mixer, or at the **IF** portion of the system can also be advantageous. For instance, for **RF** channels with a relatively narrow bandwidth, one can reduce the operating frequency of the phase shifter by placing it in the **IF** section of the electronics. Approaches like these can reduce the size and cost.

3.2.1 Commercial developments in low-cost phase shifters

A number of alternatives to the relatively expensive fabrication of hardware phase shift in MMIC are possible, and several are in the early stages of commercialization. Here we will present one specific example, and list several other possibilities. All of these have significant technical promise. Their usefulness for Naval applications will depend not only on their technical performance, but also on whether a robust commercial base for their continued development and application is established.

One example of early commercial development is based on developments in thin-film processing of ferroelectric materials [10]. One can make a time-delay phase shifter by constructing a strip line with regularly spaced varactors made from a thin-film ferroelectric such as Barium Strontium Titanate (BST). The capacitance of a BST varactor varies with applied dc voltage as a consequence of the symmetry of the Ti ion in its crystal structure. The Ti ion is moved off-center in a cell of the lattice by an applied electric field,

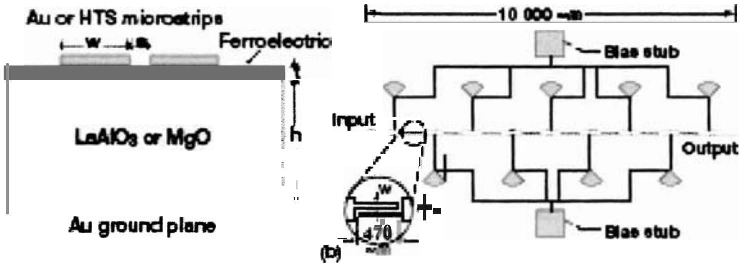
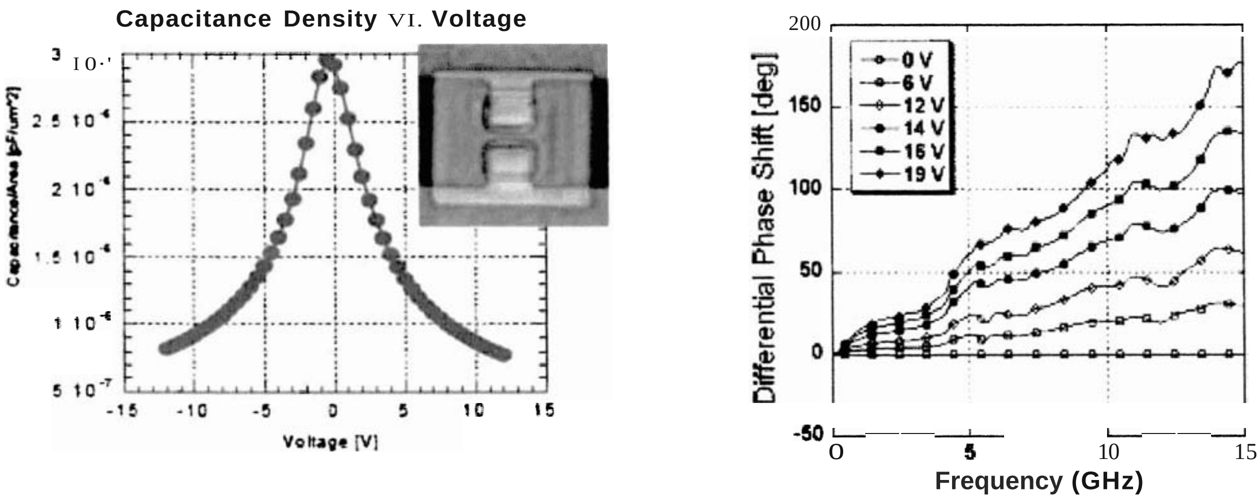


Figure 5: Strip-line design of a ferroelectric thin film phase shifter, from Reference [11]. The structure is a periodically-loaded transmission-line that can be treated as a synthetic transmission line with a voltage-variable phase-velocity.

changing the dielectric constant. This variation occurs very quickly, because it is controlled by the rapid changes in the Ti ion position.

A phase shifter with time-delay characteristics can be constructed by placing BST varactors along the length of a strip line as illustrated in Figure 5 [11]; the time delay is controlled by the voltage across the varactors. This occurs because the phase velocity of an electromagnetic pulse $v_{ph} = 1/(L_o C_o)^{1/2}$ is determined by the product of the inductance L_o and capacitance C_o per unit length of the strip line. Controlled by their voltage, the BST varactors can change the phase velocity, and thus the time delay of the strip line, by varying C_o. An insertion loss $<$ 3 dB is observed at 10 GHz for the BST-based device, however in general the RF losses are higher than desirable, and work in alternative materials is more promising [12].

The ability to make a compact time-delay phase shifter that is controlled by moderate voltages is attractive for low-cost applications. Work is continuing in optimizing the properties of ferroelectric-based phase shifters, as illustrated in Figure 6. Figure 7 presents a comparison of different varactor technologies including GaAs and MEMS. The potential of such alternative devices for low-cost components in phased array architectures needs serious



(a) **(b)**

Figure 6: (a) Insertion and return losses and (b) differential phase shifts with applied dc bias of BZN phase shifter (BZN film thickness 160 nm). For frequencies well below the Bragg frequency (determined by the periodic structure of the line), the phase shift increases linearly with frequency. At the design frequency of 15 GHz, the highest insertion loss (at zero bias) is 3.5 dB, and the maximum differential phase-shift is 175°, giving a 50 dB figure of merit at 15 GHz. From Ref. [12].

evaluation in terms of performance, integration costs, and long-term commercial stability. Alternative designs in RF-CMOS are discussed in Section 4.

3.3 Amplifers

If one were to take the concept of multi-function arrays to its extreme, e.g. supporting any and all functions on any single aperture, it would be easy to find problems with possible implementation. Of course, rational systems design will certainly lead to partitioning of functions among subsets of apertures. As noted in Section 2, the modernization program will support a substantial number of relatively low-power functions, which can be grouped without the amplifier linearily and thermal management issues

A specific example is the development of an efficient, high-power GaN HFET (Nitronex NPT 21120) for basestation amplifier applications as described in reference [18]. The devices were grown on fioatzone silicon (111) substrates. In order to make the amplifier more efficient, the drain voltage V_{dd} was adjusted with the envelope of the input signal. This amplifier achieved power added efficiency PAE = 50.7% with an average output power P_o = 37.2 W at 2.14 GHz.

The continuing commercial development of GaN devices has clearly benefited from early R&D investment by the military. The growing applications now pose the potential to allow military developments to benefit from the cost reductions inherent in higher-volume production.

3.4 Findings

Front-end architecture, e.g., the transmit-receive module (TRM) is a major cost-driver in active phased-array systems. Demanding broad functionality in a custom one-size-fits-all TRM is unlikely to be cost effective. Designs based on parallel low-cost, high-performance units optimized for different functional requirements are likely to yield the most cost effective solutions in the short term for all but the most demanding (e.g., AAW) applications. Because of the increased commercial applications for III-V semiconductor devices, there is an expanding base of COTS components that can and should be utilized in this development.

Commercially driven developments offering lower cost alternatives to existing phase-shifters are in the engineering design and test stage, in particular, MEMS phase shifters and ferroelectric strip-line phase shifters. Market-driven cost and performance analysis should be an important factor in as-

sessing their viability for active phased array systems vs. existing MMIC and RF-CMOS alternatives (next section).

Continuing development of wide-bandgap semiconductor devices for high-power RF applications is increasingly supported by commercial drivers. Exploiting the resulting economies of scale is essential to managing future costs in Naval active phased arrays systems.

4 RF-CMOS AND DIGITAL PROCESSING

4.1 RF-CMOS Capabilities and Notional Design

4.1.1 Fast semiconductor electronics

The major impediment to the widespread use of phased-array apertures has been the cost of the traditional architectures, which in turn have been limited by material costs and technical capabilities in digital processing and high-speed electronics. The increasing speed of CMOS, military R&D investment in alternative electronic materials, and market pressures due to commercial development of microwave communication and wireless networks, have resulted in new capabilities that will substantively alter the cost/performance issues governing phased array development.

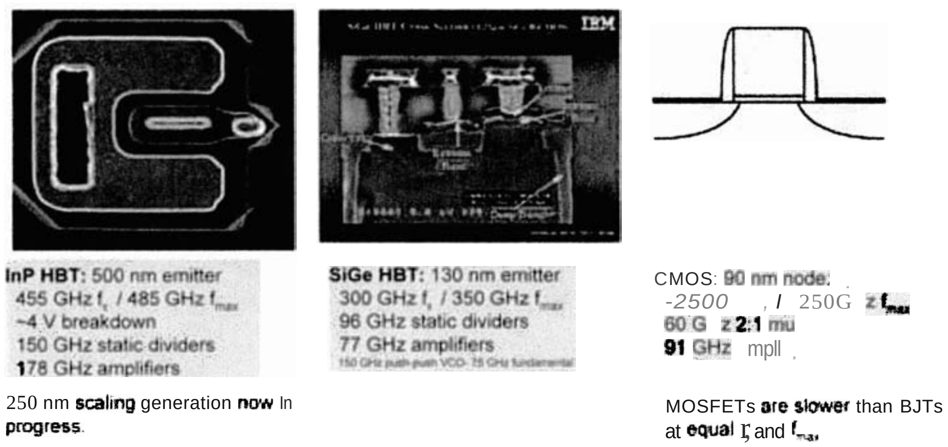


Figure 9: Integrated circuit capabilities of three semiconductor processes based on different materials, indium phosphide (InP), silicon germanium (SiGe) and silicon/silicon dioxide (CMOS). Note the different fabrication size scales for the different processes. Figure courtesy of M. Rodwell, DCSB.

Capabilities in fast semiconductor materials are illustrated in Figure 9. Speed alone does not suffice to define the optimal technology base for applications. In addition, feasibility and cost of processing complex circuits and noise figures are important criteria. Some of these trade-offs were evaluated in a perspicacious 1999 analysis for the most promising high-speed semiconductor technologies for astronomical phased arrays at 0.2 – 2 GHz. The authors predicted that SiGe CMOS would be the optimum choice, and in fact we now see that high end products and technology demonstrations based on fast RF-CMOS are being performed in SiGe Bi-CMOS. However, even as CMOS continues to move to smaller size scales, improvements in other fast materials are occurring, and the cost-benefit issues will need to be re-evaluated on a rolling basis.

In continuing evaluation of electronic materials it should be noted that in any physics analysis III-V devices have superior high-frequency performance, at the same fabrication length scale, due to intrinsically higher mobility. Furthermore, III-V devices are fabricated on semi-insulating substrates vs. the conducting substrates of standard CMOS fabrication. As a result, the standard III-V architecture is intrinsically less lossy, which is important in design of the passive structures (e.g., spiral inductors) needed in RF circuits. The material difficulties (and resulting low yields and thus high costs) in III-V processing, and the commercial investment in decreasing the fabrication length scale for Si CMOS and SiGe Bi-CMOS provide the counterbalancing effects in cost and performance.

In terms of practical performance, a properly designed CMOS or SiGe LNA ultimately should not be any worse than a GaAs LNA in terms of noise figure. The differences in performance between GaAs and Si or SiGe are now small enough at the device level that parasitics (particularly intercon-

nect losses, both on-chip and for the antenna feedline) dominate, making the overall LNA NF values differ much less. In comparing CMOS to the others, it's often useful to point out that if the spurious-free dynamic range (SFDR) is important, then CMOS actually differs insignificantly from bipolars (short channel effects behave much like built-in broadband degeneration). Indeed, CMOS usually outperforms bipolars in this regard, for equal power dissipation. Finally, it's important to emphasize that complementary mixers can exhibit much less 1/f noise than those made with n-FETs only.

4.1.2 Receive module design considerations

Evolution of TRM design into new materials systems requires rethinking the design choices that are now standard in phased array technology. Present design choices are historically linked to the analog designs of radio, then radar dishes (Figure 1), through the passive phased array designs illustrated in Figure 2, and the active array design of Figures 3 and 4. With the advent of new capabilities available in integrated circuits, every aspect of the phased array architecture can be reconsidered. In particular, not only the device implementation for each processing step should be considered, but additionally the logic governing the sequence of steps. The traditional sequence roughly follows the order:

Receive

Low Noise Amplification →

Phase Shift →

Beam forming (power combining)→

Down-conversion (mixing with LO) →

Intermediate processing→

Transmission to central processor

Transmit

Waveform generation in central processor →

Beam splitting (power splitting)→

Amplification →

Up-conversion (mixing with LO) →

Phase Shift →

Power Amplification →

Transmission

Based on the fabrication costs of various elements in CMOS, substantial changes in front end architecture can be considered. For instance, a model architecture for a (possibly integrated) receive module combines the LNA, the first mixer(s), and the local oscillator (LO) for the first mixer(s) as shown in Figure 10. Narrow-band beam steering (a phase shift) can be done almost

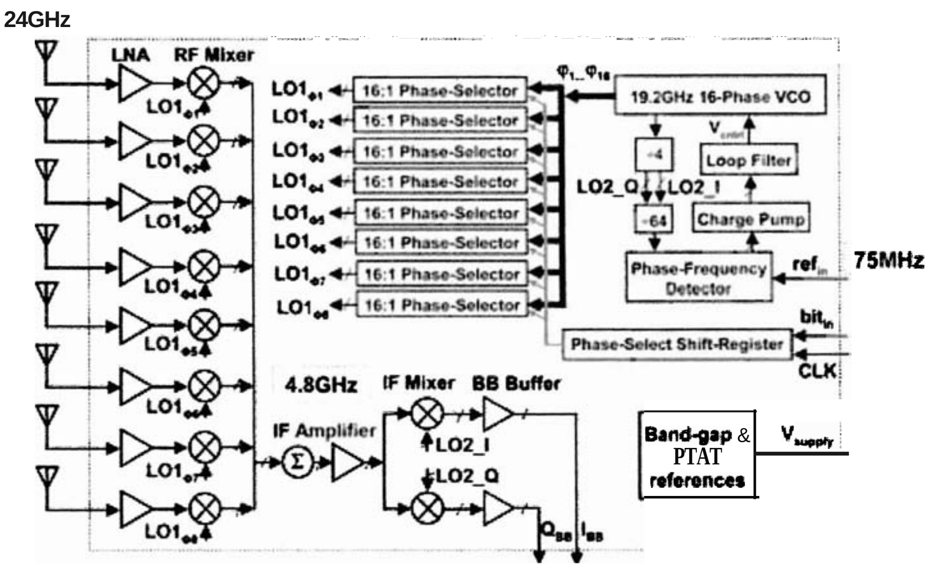


Figure 10: Diagram of the receive architecture for a 16-element phased array. Design has been implemented for a 24GHz system using SiGe biCMOS, and beam forming with 4 elements has been demonstrated. From Reference [19].

for free by phase shifting the La (relative to a reference phase) before mixing. This necessitates a separate mixer and La phase shifter (and possibly LO) for each beam being formed. However, these mixers are cheap (when integrated), so supporting 8-16 simultaneous beams (channels) is not difficult. Note that the phase shift setting for each element is digital, a digital input to a phase interpolator in the La phased lock loop (PLL) feedback 100p.2 Also note that, unlike traditional designs, here the down conversion occurs BEFORE beam forming. The full phase/amplitude information of the individual waveforms for all the antenna elements therefore can be available at intermediate frequency (IF) for whatever level of digital processing that can be supported in the front end circuitry.

In implementing the down-conversion, one possibility would be to use a commutating mixer to mix the signal down from RF to baseband. This mixer can easily be implemented in CMOS and can double as the sample and hold for the subsequent ADC. Using a commutating mixer, however, will fold the spectrum over many times, and so depends on filtering out the desired channel before the mixer. Using Gilbert-cell mixers (or any non-sampling mixer) enables the use of **IF** filtering.

After the mixers we then have an **IF** signal (I and Q) for each channel of each array element. All of the element signals for each channel need to be summed to form the beams. This summation can be done in the digital or analog domain. Performing the sum in the digital domain reduces noise and

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2There is nothing fundamental about 200 MHz bandwidth. It is just used here to provide a concrete example. Converters are available commercially at up to 3 Gs/s, so that 1.5 GHz channels could be implemented, with better performance available for custom-designed converters. Another limit on the width of individual channels will be the issues of amplifier linearity and true-time delay vs. phase shift in implementing steering.

adds flexibility. Performing the sum using analog signals, as shown in Figure 6, reduces the number of ADCs needed dramatically but also reduces flexibility – and does not allow the wide-band beam forming described below.

While analog summing requires fewer ADCs, it requires better ADCs. Suppose we need 12 bits of output precision [20] for net output of each channel of the array. With digital summing, we could convert at each element using 7 bits of precision and then sum over 1024 elements to give an output signal with 12-bits of precision. As long as the quantization noise is uncorrelated (which can be arranged with dither), it will sum as $N^{1/2}$ over the N elements. With analog summing we would need a single 12-bit converter. If the highest bandwidth for a single signal were 200MHz,[3] we would need 400 m sample/s converters for I and Q at Nyquist sampling rate. Sampling at this rate is feasible at either 7 or 12 bits, but a lot less technically demanding at 7 bits, as suggested by Figure 11. Further design considerations would have to be addressed if significant jamming issues, or weak/strong signal issues cannot be handled within 7 bit digitization.

With the ADC at the elements and subsequent digital summing, the output from (and phase control input to) the elements is entirely digital. With a 400 MHz 12-bit channel signal, the output bandwidth can comfortably be transmitted on a single twisted pair up to 5-10m. (Note that the PCI-Express 2.0 standard uses 5.0Gb/s per pair). For a large array

[3] If improperly implemented PLLs can add phase noise. Thus careful design would be needed if this particular aspect of the approach were to be implemented for Naval systems. Specifically, the output phase noise depends on the input phase noise, the PLL loop bandwidth, the phase noise characteristics of the PLL's internal VCO, and the implementation details of the feedback divider. Within the loop bandwidth, the output phase noise will track that of the input (within a multiplication factor corresponding to the divide ratio), so if the input is clean, the output will be (or can be made to be), also. Outside of the loop bandwidth, the inherent phase noise of the PLL's VCO will dominate. This requires care in the VCO design as well.

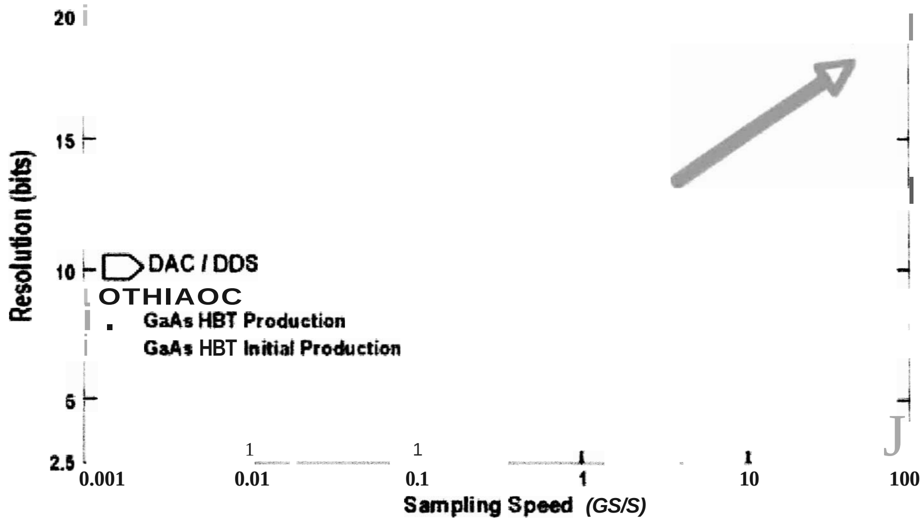


Figure 11: State of the art ADC resolution vs. sampling rate circa 2004. The red arrow indicates the path of technology development, e.g., toward higher speeds and resolution. Figure courtesy of M. Rodwell, UCSB.

with 10^3 - 10^4 elements, these serial channels would be combined digitally in a module co-located with the array (see also Section 4.2). If sufficient compute-power could be located at the array, this module would perform the wide-band beam steering. Alternatively, as discussed later, more limited computation involving local calibration and intermodulation correction could be performed at this level, with transmission of the summed signals to the central processor.

Performing all I/O from the elements digitally greatly simplifies the problem of connecting up the arrays (and reduces the cost of this connection). The only analog signal that would leave a receive module in this design is the single phase reference signal that is broadcast to all of the elements for each channel. This signal needs to either have a matched delay to each element - or the phase shift of this signal to each element needs to be characterized so it can be offset with the digital phase adjustment. The signal out of the receive module and the phase control signals to the receive module are

entirely digital and hence insensitive to delay mismatch and small amplitude variations.

4.1.3 Transmit module design considerations

Broadly speaking, the transmit module is the same as for the receiver module but in reverse. Using the same line of argument as above, data and phase information can be distributed in digital form to the transmit modules in the array. The transmit modules then would each contain an independent DAC, LO, mixer, and a power amplifier. Sequentially, the DAC conversion from digital signal to analog will occur first, followed by signal up-conversion via the LO and mixer. If multiple beams are to be transmitted, then the multiple channels would be combined at the analog stage, and the summed signal would be passed through the power amplifier and coupled to the antenna element.

4.1.4 Active phase correction

In the system design discussed above, the phase reference signal can be internally auto-calibrated for the individual transmit and receive arrays, eliminating the need for delay matching. This can be accomplished by connecting phase signals from adjacent elements together. Specifically, this would involve distributing the locally generated oscillator signal to neighboring modules, each of which will then input this signal to a phase comparator. The signal is an RF signal, but a very low power RF signal and requires no modulation. Signals from mutiple neighbors can be selected with a multiplexer, or using multiple phase comparators. During a calibration phase, each ele-

ment would in this way measure the relative phase between itself and its four neighbors, and calculate a phase correction to bring them in phase. After a few iterations of this phase correction, all modules would converge on a common zero phase without the need for any matched-length cabling of the reference signal. The level of phase calibration needed is directly related to the side lobe level that can be tolerated. For example, quantization error at 6 bits, or one degree of phase error, yields side lobes at -32dB.

4.1.5 Issues for Wide-Band Systems

For wide-band beam steering, the issue of true-time delay vs. frequency-dependent phase shift (see Section 3.2) must be addressed. If different frequencies correspond to independently steered beams, than implementing individual phase shifts is a natural choice. For an intrinsically broad-band signal, where all the frequencies would be steered in a single beam, true time delay would be preferable. For either of these approaches, it is better to convert the signal to digital first and do all of the steering in the digital domain. The same organization described above would be used, but with the phase shift setting for each element set to zero. The IF (or baseband) signal would then be digitized and either delayed-requiring an interpolation filter to get sub sample-time resolution – or split into sub-bands - with a polyphase filter bank.

4.2 **Full** Digital vs. Intermediate Beam Forming

Pure "software controlled RF systems," that is full digital control of the waveform applied to each element of the array, for multi-function phased

array is a compelling concept. It would allow the direct generation of simultaneous beams with adaptive beam shaping, overcome many limitations of analog hardware for wideband signals, and allow new applications such as pulse-chasing receive beams [21]. The implementation is illustrated schematically in Figure 12. It is clear that digital conversion speeds (ADC and DAC) are an important issue in the feasibility of such a design for use at high frequencies. Less obvious, but equally challenging, is the issue of data transmission from the front end to the CPU (labeled as COTS processor in Figure 12).

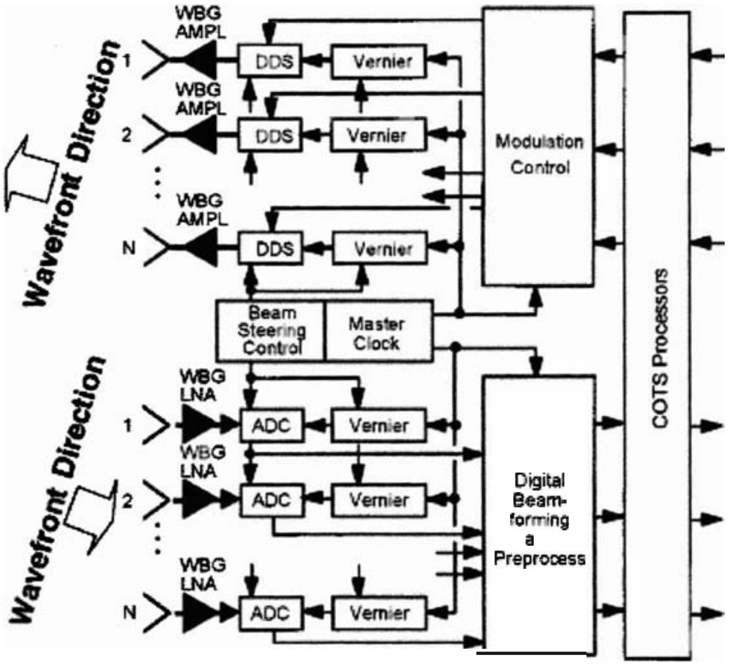


Figure 12: Block diagram of a notional full digital phased array architecture. The power amplifiers (on transmit) and low-noise amplifiers (on receive) would be the only analog circuits at the aperture. Figure from Reference [89].

4.2.1 Alternative approaches to fast ADC and DAC

Obtaining high resolution in signal conversion will be important when mixed signals of different strength are present, or when signals are obscured by noise or jamming. As noted above, obtaining higher resolution is increasingly difficult at high conversion rates. A further complication arises when digital to analog conversion (DAC) is needed at a very high data rate and at high power, e.g., for high power transmissions from a wideband aperture. **In** addition to the approach of developing faster transistors, there are a number of of clever design approaches that can be used in obtaining higher conversion rate and resolution [22].

The $\Delta\Sigma$ DAC design comes from a desire to provide higher resolution in the DAC output than is allowed by the number of bits in the converter. For example one may have an 8-bit DAC that has sufficient accuracy, but 16-bit resolution is needed. This is, for example, the case in using the digital bit stream from a compact disc (CD) to provide analog signals to drive analog amplifiers and/or speakers in music audio systems. The trick is to run the DAC at a much faster speed than the maximum frequency response of the system one is driving with the analog output. **In** this situation the DAC can be dithered between low resolution levels to produce a higher resolution after averaging the DAC output. A simplified $\Delta\Sigma$ DAC block diagram is shown in Figure 13. **In** this concept the feedback loop is run many times faster than the desired input value changes, so that the output shown in Figure 14 changes many times between changes in the desired value as provided at the input. However, these many changes between the low resolution values of the DAC output allow the average (over the time scale of the input signal changes) to become much closer to the desired value than the resolution of

the DAC. This operation is shown in Figure 14. The desired value in this example is 1.62V. Note how the DAC dithers between the allowed values of 1 and 2 to reach 1.62 on average.

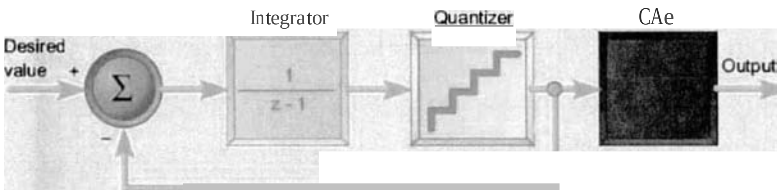


Figure 13: Simplified block diagram of a ΔΣ DAC. In a basic ΔΣ modulator, the desired output value is applied to the summing junction along with the actual output. This delta value is added to the integrator value (sum) and applied to the quantizer. After Wescott [24].

The ΔΣ DAC concept works well and is commonly used, especially in producing analog audio from digitally coded sound, e.g., CD players and iPods. However, the control loop rate is many times higher, typically 64 times faster than the input digital data stream. The control loop cycle rate being 64 times faster than the input rate is referred to as 64 times oversampling. This is no problem at audio frequencies where the maximum output frequency does not exceed about 20 kHz and the Nyquist sampling rate is about 40 kHz. The sampling rate of the control loop is then 64 x 40 kHz or a few MHz - easily done in inexpensive electronics. The output of such a DAC is very linear and yields very high quality audio with very little intermodulation distortion. All the switching at high frequencies produces noise, but at frequencies much higher than the operating band. A low-pass filter typically follows the ΔΣ DAC output to eliminate the high frequency noise. Schreier and Themes [23] provide a full discussion of ΔΣ DAC's and ADCs as well as applications.

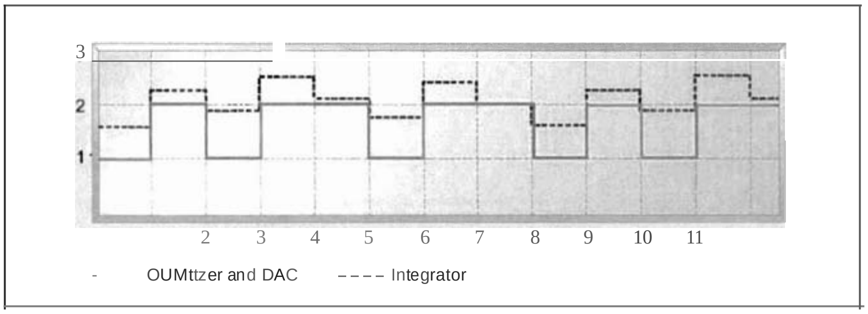
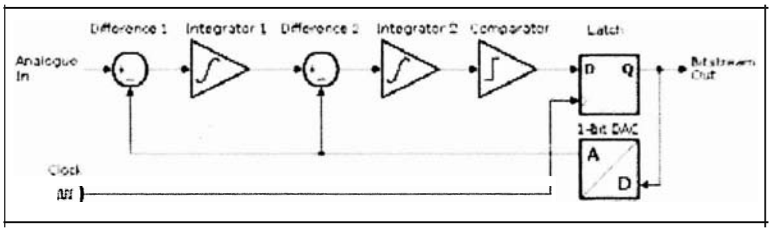


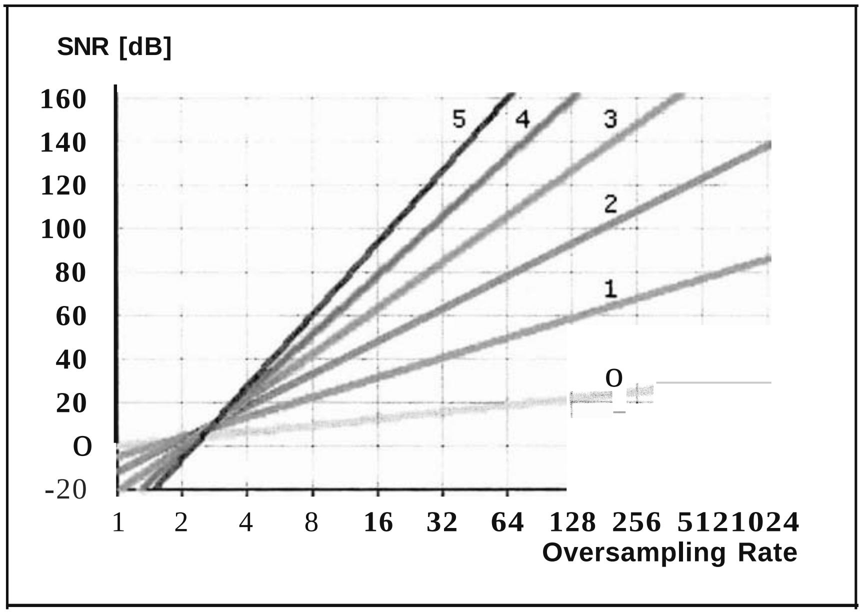
Figure 14: Output of $\Delta\Sigma$ DAC showing how the average value of the DAC output approaches the desired value when the average is taken over the time scale of the input (desired) signal changes, i.e., over many cycles of the control loop. The desired output is 1.625, and the DAC dithers between 1 and 2 to reach this average value. High and low portions of the cycle are spread out as much as possible. This means that whatever the DAC is driving should have a much slower response to the variations in the DAC output than the control loop cycle time. After Wescott [24].

It is tempting to apply these techniques to high-frequency data conversion. However, if the maximum frequency of the signal we are trying to produce is say 10 GHz, oversampling of 64 X would require a control·loop operating at 640 GHz - a very difficult proposition at present (see Figure 11). However, in principle, the oversampling rate and the conversion noise can be held down by using a higher order $\Delta\Sigma$ scheme, such as shown in Figure 15. We point out that orders higher than two can not be done by simply adding more stages as in Figure 15a, but use low pass filters. Fifth order $\Delta\Sigma$ schemes are commonly used in audio frequency equipment. In Figure 15b we show an estimate of the impact of both the order of the DS scheme and the oversampling factor. The point is that higher order DS schemes allow one to achieve respectable signal to noise ratios with oversampling rates that are much lower than the typical factor of 64.

Another approach to increasing resolution is the use of interleaved conversion structure, which has had substantial success in the development



(a)



(b)

Figure 15: a.) Schematic block diagram of a 2nd order $\Delta\Sigma$ modulator. Note the two feedback loops. Although an ADC is shown, the DAC is analogous. After Beis [25]. b.) Conversion signal to noise ratio as a function of oversampling rate and order of the $\Delta\Sigma$ structure. After Beis [25].

of wideband, software defined test instruments. Here, a number of low-resolution converters are run in parallel with a calibrated time offset and a common offset. If requirements on the spectral purity of the signal conversion are low, simple implementation of this approach can be quite successful. However, in phased-array techniques where signal identification occurs in the frequency range, careful calibration and correction of the parallel converters is needed to obtain the spurious-free dynamics range (SPDR) needed. Rodwell and coworkers [22] accomplished this using finite-impulse-response (FIR) filters, with the results shown in Figure 16.

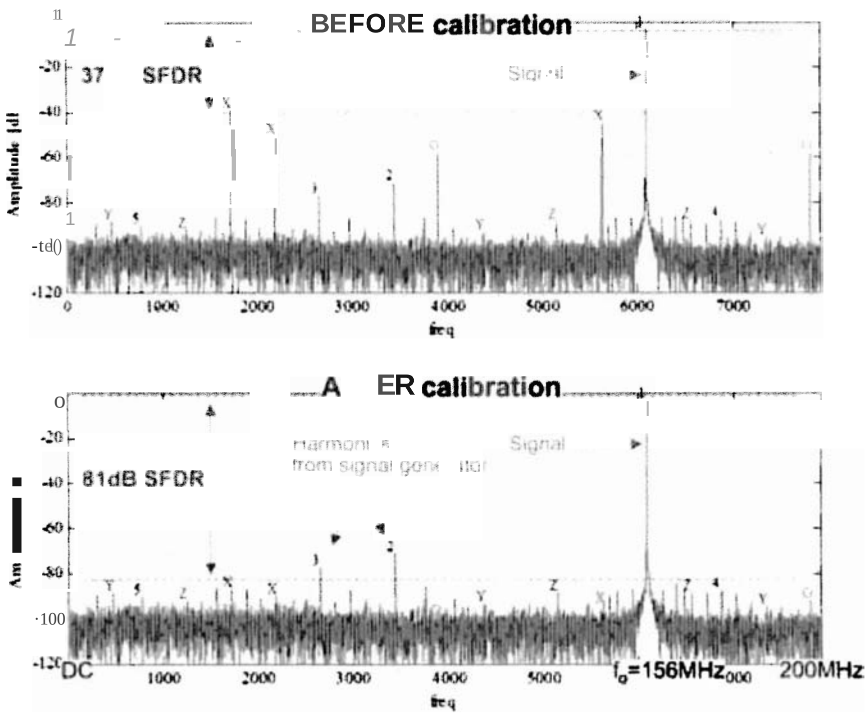


Figure 16: Effects of calibration and correction on time-interleaved analog-to-digital conversion on a prototype 14 bit 400 MSPS ADC. Peaks labeled X,Y,Z are respectively 1st-order, 2nd-order, and 3rd-order gain/time spurs.

These and related advanced architectures involve substantial increases in design complexity, and in many cases also involve extensive digital processing for correction and calibration. Substantial research investment in such

approaches should be balanced with on-going evaluation of the trade-offs in performance and cost.

4.2.2 Data Transmission

In the design architecture presented in Figure 10, the digital signal generated at each element of the receive module was combined at the array and sent as an I-Q digital signal (carried on the array) down to the radio room as shown schematically in Figure 17. The major factor in deciding to sum signals at the array, and thus loose the full phase information, is the cost of transmitting all the individual receive module signals to the control room.

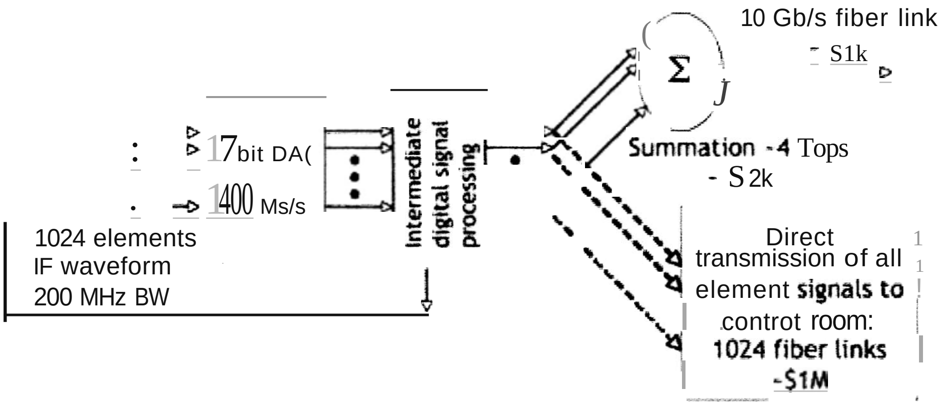


Figure 17: Schematic illustration of two limiting choices for transmitting array data to the central processor in a receive array. In one case, beam forming is done prior to transmission. In the other case all the individual waveforms are transmitted in parallel and beam forming is done in the CPU. The issues for a transmit array are similar, with the summation replaced by power splitting.

The cost trade-offs can be illustrated with some simple estimates, as follows. Combining the signals from the elements for a narrow-band beam costs a complex multiply-add per element/sample. For a 10^3 element array at 200 MHz we have – 8 ops/el samp x 10^3 elements x4 x 10^8 samples/s =

1.6×10^{12} - or a requirement of just under 4 Tops/s. This can be accomplished with 20 high-performance processing chips (e.g., Stream Processors) for about *$2K*. The resulting single channel stream of I and Q samples is then 12 b x2 x 4 x 10^8 = 9.6 Cb/s, which can easily be sent down to the radio room on a single lOCb/s fiber at a cost of about *$1K*. Thus the total cost is about *$2K* of processing and *$IK* of communication.

On the other hand, if all of the element feeds are transmitted separately, then each will require a *$IK* optical communication link for a total cost of $IM (for 10^3 elements). These cost numbers are rough estimates to within a factor of ~ 2, but the result is still clear: it is much more cost-effective to combine elements at the array and have a single output per channel than to try to provide an output per element x channel. Power considerations will also favor this solution, since the power dissipation of 10 processing chips, 100-200W, is much less than the power of 10^3 fiber transceivers–at least lkW.

The analysis of the two limiting model cases presented above is based on the use of the present 10 Cb/s telecommunications standards for serial links. Higher data-rate standards, with 40 - 50 Cb/s data rates [26, 27] are nearing commercialization. Commercial pressures should bring down costs and improve performance with respect to integration and power dissipation, making this an attractive technology for use in multi-function array architectures. Whether this will significantly change the cost balance for beam forming at the array vs. in the central processor remains to be seen.

Note that in this model for the architecture, the bulk of the beamforming algorithm - the part that computes the weights (narrow band) or delays (wide-band) remains in the radio room and sends the weights and delays up to the array via fiber. For some algorithms, in particular those

responsible for dynamic error correction, calibration and predistortion (see following section), a part of the algorithm may need to run at the array - to compute on the raw data before combining. This will increase the computational demand at the array, leading to designs with larger signal processing capabilities near the front end. Given decreasing costs for signal processors, this is likely to be a favorable design choice. It has the additional benefit of increased redundancy in the operation of multiple arrays.

4.3 Commercial Drivers

The ability to steer multiple different signals at different frequencies, to move rapidly between signals at different frequencies, or to deal with steering a single extremely wide band signal are all technical needs (see Table 1 in Section 2) of the Integrated Topsides program. Addressing these issues with simple extrapolation of traditional, analog active phased array architectures, as discussed in Section 3, leads to serious problems due to amplifier linearity and efficiency, cost, and concerns about cross-talk and intermodulation for parallel analog signals. Alternative architectures, based on the digital signal processing capabilities of CMOS, as in the examples in Section 4.1, offer promising new approaches to some of these issues. These include lower cost for front-end electronics that allow parallel processing of different frequency bands, and the possibility of digitization early in the receive (or late in the transmit) path. Serious remaining issues are the bandwidth of the signal to be handled, and the difficulties of fast analog/digital conversion when a large dynamic range is needed. Increasing pressures on availability of commercial spectrum are leading to substantial commercial investments in using broadband signals. The commercial developments of particular relevance are the

802.11 standards for local area networks (2.4 and 5 GHz bands), the recent FCC decision to allow intentional low-power emissions in the 3.1- 10.6 GHz region, and the commercial developments in multiple -input/multiple-output (MIMO) for reducing multipath and increasing transmission efficiency. The relationship of these subjects to wideband phased array architectures is discussed in the following section. An example of an emerging commercial technology using SiGe BiCMOS is then presented.

4.3.1 Multi-band, Ultra-Wide Band and Multiple-InputjMultiple-Output

The orthogonal frequency division multiplexing (OFDM) used in the 802.11 standard provides an example of information transmission using multiple parallel frequencies. Specifically, sampling rate is reduced in OFDM by dividing the data stream into a number of lower-rate channels that are then stacked in frequency. In 802.11a, the wide frequency band from 5.15 to 5.825 GHz is subdivided into channels of 20 MHz width, each of which can carry a power from 40mW to 800mW (compare Table 1 in Section 2). Each channel is then further subdivided into 300 kHz bandwidth sub-carriers that transmit a coded stream of data. The 300 kHz band width allows inexpensive ADC/DAC and digital processing. Implementation has been demonstrated in RF CMOS, for instance in the Atheros AR5006 [28], which has a data rate to 108 Mb/s. The commercial developments of OFDM demonstrate the power of using CMOS digital signal processing (DSP) to generate or decode hundreds of separate channels simultaneously. A conceptual architecture for adaptation of this concept for high-data-rate satellite communications in 4-channels at 622 Mb/s is shown in Figure 18 [29]. This type of architecture could clearly be used to in both transmit and receive for signals covering a

wide range of frequencies. While issues of co-channel and adjacent channel interference must always be addressed, these are well-recognized problems that can be predicted with existing engineering codes.

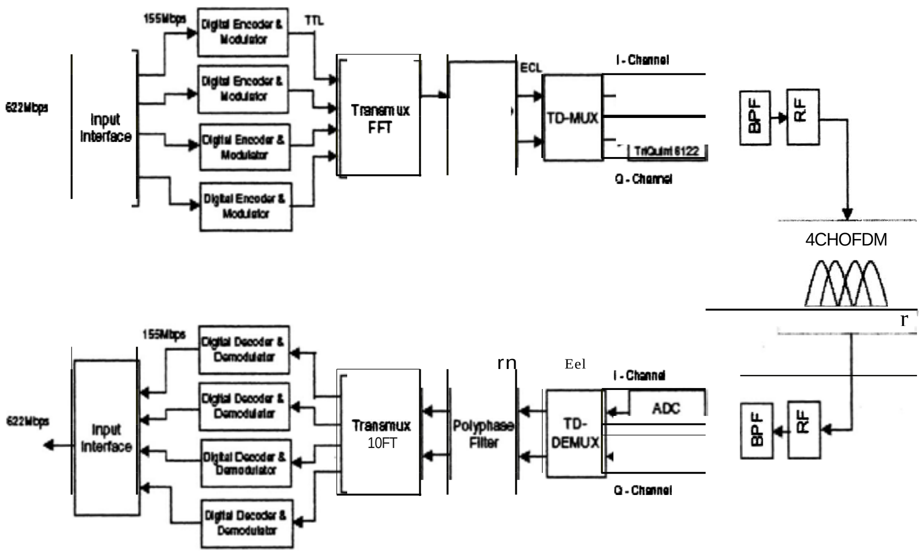


Figure 18: Architecture for a four-channel fast communications system using OFDM concepts. The basic OFDM waveform is constructed by dividing an incoming data stream into four channels, encoded and modulated. On receive, each channel undergoes discrete Fourier transform (DFT), creating a digital label for each complex component of the signal. these are further translated into amplitudes in the polyphase filter. Summation of the component signals occurs in the (multiplexier) MUX. On transmit, noise corrupts the amplitudes extracted from the ADC and DEMUX and the polyphase filtering must be designed to match the ADC process.

Expanding bandwidth even further, as a way of more efficiently exploiting spectrum, is a major goal of Ultra-Wide Band technology [30, 31]. The FCC has recently given permission for low-power transmissions (e.g., below the noise limit of existing allocations) in the 3.1- 10.6 GHz frequency range. Utilization of the low-power transmissions involves switching data-streams among different 500 MHz wide bands so that on average, no individual band exceeds the power-limit, and transmit and receive signal processing is lim-

ited to the smaller bandwidth. Switching among various 500 MHz bands could be accomplished by switching the local-oscillator frequency (LO) used for the up-conversion. As discussed in Section 4.1.2, this could be accomplished relatively easily in CMOS by the use of variants of Gilbert cell mixers, and phased lock loops to generate the different LOs.[4] Thus, multiple 500 MHz bands could be combined in the same transmit/receive module (TRM), subject to the bandwidth of the LNA or PA. Development of ultra-wide-bandwidth amplifiers over the 3-10 GHz range using multiple band-pass Chebyshev filters has been demonstrated in CMOS [32], and continuing development is motivated by the expected commercial applications in UWB [33]. As UWB products gain market share, the related technology developments can increasingly be exploited for Integrated Topsides.

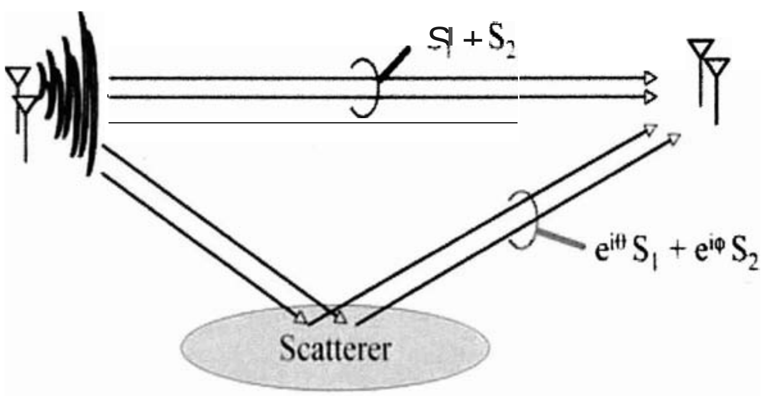


Figure 19: Illustration of MIMO concepts a) signal propagation with multipath interference can be exploited to improve signal transmission in configuration with multiple antennas, figure courtesy of S. Simon of Lucent [34].

4It is significant to note that complementary oscillators can be made to exhibit superior phase noise properties (for a given resonator Q and power budget). CMOS has this property, whereas bipolar technologies ordinarily do not provide true complementary devices. So, for the VCO inside the PLL, a CMOS implementation would be attractive.

The use of multiple antennas on both transmit and receive has gained increasing attention, first because of its potential increased transmission rates by exploiting multipath [34], as illustrated in Figure 19. Substantive signal processing is used to achieve the gains. More straightforward benefits of MIMO can be achieved using the concepts of beam forming [35]. Even rudimentary beam-steering is of substantive commercial interest because of the potential to increase directional power transmission, and reduce interference in RF-crowded environments. Increasing sophistication of beam-forming applications is expected in the future in commercial RF products, for instance in wireless connections for lap-top computers and in vehicular radar. This in turn will provide market drivers for low-cost beam-forming architectures and approaches. It is quite reasonable to expect that the oversampling techniques that work well at the lower frequencies relevant to digital image processing [36] and Ultrasound Imaging [37] will be evaluated for their potential commercial uses in RF front ends [38], including the demanding implementation for beam forming [see Appendix D].

4.3.2 Commercial RF-CMOS

While the examples of RF-CMOS discussed in Section 4.2 were in the research and early development stage, it is important to realize that fieldable products for high-end applications have already been developed. An example is a program by the Intelligence Technology Innovation Center (ITIC) to develop highly integrated broad-band RFICs for reconfigurable radios [39]. In a one-year development program, they were able to bring a successful product to completion. The circuit consists of four of the units illustrated in Figure 20, and has a fabrication cost of $\sim$ $200 per chip. With larger-volume production, it is likely to be possible to bring per-chip cost down to $50.

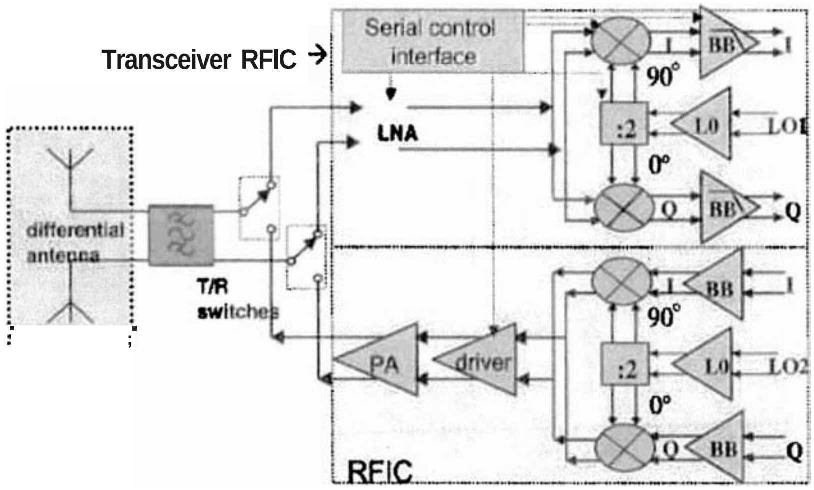


Figure 20: Receiver plus transceiver unit for operation over 100 MHz to 3GHz receive and 1.1-2.2 GHz on transmit. Fabricated in 0.18μm SiGe BiCMOS. Receiver includes LNA with up to 25 dB gain, two doubly balanced mixers, with a combined noise figure <1.8 dB. Transceiver includes a 20 dBm PA with output power of 100 mWand a transmit gain of 40 dB. Figure courtesy of ITIC [39].

The full circuit includes both RF (shown in Figure 20) and baseband circuits. The base band gain control is applied with variable gain amplifiers (20 dB gain, BB in Figure 20) and a WCDMA filter (46 dB, not shown) controlled by a digital stream through a serial interface circuit. The bandwidth can be varied from 100 MHz to 300 MHz, and frequency tuning is controlled by DSP on the output signal with feedback to the filter tuning control inputs. The output of the receiver (input to the transceiver) is converted at 8 bits resolution with on-board ADC (DAC). The specifications for one (of the three available) receiver designs is shown in Table 2.

The circuits were fabricated in 0.18 μm SiGe BiCMOS. The choice of the materials and design format were dictated by the requirement for broad-band performance and low noise (see Section 4.1). The present design is readily adaptable to frequencies up to 10 GHz, and the transmit power could rea-

Table 2: Performance specifications for the receive channel of the RFIC circuit. Figure courtesy of ITIC [39].

Characteristic	Typical Specification
Overall receiver specifications	
Operating range RFIC	100 – 6,000 MHz
Overall receiver noise figure	<3.0 dB up to 3 GHz <4.5 dB up to 6 GHz
Overall receiver IIP3	-12dBm at nominal VGA gain
Overall receiver IIP2 (typical at 1 GHz)	+25 dBm
Overall receiver P_{1dB}	-23 dBm at low baseband gain
Dynamic Range (no gain variation)	60 dB (CDMA)
Baseband Gain Variation	65 dB
Programmable IF bandwidth	1.1 - 3.3 MHz
Amplitude balance quadrature mixer	<1 dB
Phase balance quadrature mixer	< 1 deg
Receive sideband suppression	-40 dBc
LO to RF leakage	< -85 dBm
Overall Receiver RFIC power consumption	250 mW
Power sources required	+3.3 V
Overall receiver chip size	2.7x2.3 mm
External Oscillator specifications	
Frequency	double frequency, quadrature using divider
Minimum LO Power required	-20 dBm
Number of bits	27
Average Power Consumption	10 mW

sonably be increased to 1W per channel, appropriate for all the applications in Table 1 except AAW. Adaptation of such circuits in environments where jamming or weak/strong signal mixes are significant could be addressed by rapid time multiplexing over the frequency bands, or by implementing parallel hardware channels with static or programmable frequency filters.

4.4 Active Correction of Nonlinearity

4.4.1 Introduction

A well-known problem in RF communications is the inefficiency of highly-linear amplifiers, such as Class A amplifiers, which may have efficiencies of

only a few percent in their linear range. Efficient amplifiers, as in Class D, on the other hand, may have non-linearities that create intolerable distortion. In such cases, pre-distortion techniques are used to undo the non-linearities, supposing that some sort of knowledge of the non-linearities is available through tests such as measuring the distortion of a known signal with two or more tones. There is a long history of analog pre-distortion techniques, but with the advent of faster and cheaper computing power it becomes more and more interesting to consider digital pre-distortion, in which the operations needed to undo distortion are calculated digitally. Moreover, analog pre-distortion methods are complex and do not scale well in size as other RF components get smaller.

This is not the place to go into great detail about digital pre-distortion, but we consider some general features of intermodulation (1M) distortion in a bandpass-limited system with memory that may well be amenable to digital calculation, depending on the system bandwidth. We only consider the problem of in-band distortion, although mixing of two signals in different bands can well lead to unwanted signals in a third band (see Section 5.1.1).

There is a good deal of current interest in digital pre-distortion, with several solutions available for memoryless models, that is, models where the output $So(t)$ at a given time t is a simple polynomial function of the input $S1(t)$ at the *same* time:

$$S_O(t) = gS_I(t) + \epsilon S_I(t)^3 + \dots . \tag{4-1}$$

(Of course, many other terms may be present; we only write a cubic term, which has in-band non-linearities such as 1M distortion; a quadratic term leads to out-of-band effects. Only odd powers of $S1$ contribute to in-band non-linearities.) Algorithms for distorting the input so that the output is linear in the input usually involve a lookup table for properties of the non-

linearity, such as the strength ϵ, with some simple processing to invert Eq. (4-1). An example is [40]. Since it may happen that accurate and efficient removal of 1M distortion involves taking into account of the memory of the hardware amplifier (so that its output depends non-linearly not only on the input at output time but also, through a convolution, the input at prior times), we discuss in a very general way how one might approach digital pre-distortion in band-limited systems. The main point is that in a band-limited system one has, in effect, automatic discretization of time through Nyquist sampling, and this discretization, plus a corresponding discretization in frequency space, allowing the use of fast Fourier transforms, is suited for digital manipulation.

An interesting subclass of these in-band problems is the 1M distortion produced in a phased-array antenna broadcasting two or more signals in the same band but on different beams (see Section 5.1.1). In some cases it may be necessary to reduce the 1M distortion substantially below the expected strength of one beam's sidelobe in the main lobe of another beam. These multiple signals are usually characterized by essentially sinusoidal wave forms of about the same frequency with stable and well-known phase shifts between different beams. This raises the prospect that digital pre-distortion can make very effective use of the known signal characteristics, especially the different phase shifts. We know of no existing treatment of this problem, although it may well exist; we give a brief discussion of the general principles of phase manipulation to conclude this chapter. Here, to simplify things, we assume a memoryless model of the type of Eq. (4-1).

4.4.2 A bandpass-limited model with memory

Consider a system with bandwidth B, so that every signal that can be input to or processed within this system has the form

$$f(t) = \frac{1}{2\pi} \int_{-\Omega}^{\Omega} d\omega e^{-i\omega t} \tilde{f}(\omega) \tag{4-2}$$

where $\Omega = 2\pi B$. There is no need to demand that $f(t)$ be real, which would require $\tilde{f}^*(\omega) = \tilde{f}(-\omega)$. Instead, we take it that $f(t)$ is represented as $I + iQ$, where I and Q are the real in-phase and quadrature parts. The spectral function $\tilde{f}(\omega)$ can, for a bandpass-limited signal, be written as a Fourier sum:

$$\tilde{f}(\omega) = \sum_{-\infty}^{\infty} \tilde{f}_N \exp[\frac{2\pi i N\omega}{2\Omega}]. \tag{4-3}$$

The Nyquist sampling theorem, following from using Eq. (4-3) in Eq. (4-2), tells us that in such a case an alternative form for $f(t)$ is

$$f(t) = \sum_{-\infty}^{\infty}[\frac{\sin \Omega t}{\Omega t - \pi N}] f(t = \frac{\pi N}{\Omega}), \tag{4-4}$$

that is, the function can be exactly reconstructed by sampling at the Nyquist rate $2B$. **In** deriving this result, one uses the relation

$$\tilde{f}_N = (-)^N \frac{\pi f(t = \pi N/\Omega)}{\Omega}. \tag{4-5}$$

Consequently, the Fourier series expansion of the spectral function $\tilde{f}(\omega)$, Eq. (4-3), is a series representation of this spectral function in terms of the Nyquist-sampled time values of the function $f(t)$. A signal that vanishes identically for $t < 0$ is one whose discrete Fourier series has only terms with $N > 0$.

We use a Volterra model, which is a general representation for systems

with memory [41]. **In** Fourier space the Volterra model becomes:

$$\tilde{S}_O(\omega) = \sum_1 \left(\frac{1}{2\pi}\right)^{N-1} \int_{-\Omega}^{\Omega} \prod^{N} [d\omega_i \tilde{S}_I(\omega_i)] \tilde{H}_N(\omega_1, \ldots, \omega_N)\delta(\omega - \sum^{N} \omega_i). \quad (4\text{-}6)$$

Here $S_{(O,I)}(\omega)$ is the Fourier transform of the output (O) or input (I) signal and $H_N(\omega_1, \ldots, \omega_N)$ is the Fourier transform of the Nth Volterra kernel; the δ-function is the Dirac delta function. These non-linearities would appear to extend the bandwidth of the output to $-N\Omega < \omega < N\Omega$ for the Nth kernel, but the system, by hypothesis, is only capable of dealing with signals within the $N = 1$ band, and so we will impose this constraint on Eq. (4-6). We therefore limit our attention to in-band IM distortion, which must be reduced when a strong and a weak signal appear in the same band, as in the near-far problem.

By substition of the discrete Fourier series of the type in Eq. (4-3) for the various quantities with tildes in Eq. (4-6), it is straightforward to see that the result is a version of the original Volterra equation in the time domain, but with discrete times t_k, where k is an integer and $t_k = \pi k/\Omega$ is the Nyquist sample interval. This suggests the following strategy:'

1. Choose a maximum total time over which a bandpass-limited signal will be sampled, and denote this maximum number of samples as $\mathcal{N}$; take this to be an even integer. A longer signal in time is constructed by overlapping a number of such length-N samples, each displaced by one sample unit from the previous one.

2. Discretize frequencies as well as time, denoting the discrete frequencies as

$$\omega_j = \frac{2j\Omega}{\mathcal{N}}.$$

for integral j running from $-\mathcal{N}/2$ to $\mathcal{N}/2$. This allows the use of fast Fourier transform (FFT) technology at appropriate places.

3. Evaluate the output signal in terms of the input signal by using the Volterra model as a sum over discrete time inputs, as discussed above.

4. From the evaluation of the output in terms of the input as in steps 1-3, one can invert the results to find an operation which equalizes the original non-linear system so that, for example, a weak signal can be amplified linearly even in the presence of **1M** distortion from a strong one. There are standard techniques for doing this, for example, [42], which we need not discuss here.

Suppose that the non-linear system has only linear and cubic responses, and that the issue is the near-far problem. The third-order problem to be solved is quadratic in the strong signal and linear in the weak signal. (In such a case, the work to be done for step 4 above is a linear inversion.) To evaluate the Volterra equation requires two FFTs and is $O(N2)$, but every update step past the complete treatment of the first N samples requires only $O(N)$ updatings for every sample.

Just how computationally complex all this might be in practice is hard to tell, but it clearly requires processing power which is at least N times the Nyquist rate, and so is dependent on the signal bandwidth. **In** general, the more bandwidth one puts into a single channel, the more stressing the computational job, so it makes sense to channelize shipboard RF functions if digital signal processing is to be really useful, whether the job is pre-distortion or something else.

It is also worth noting that a good scheduler, somewhat in the style of a cell-phone base station, can substantially relieve the load of pre-distortion digital processing. Presumably time-division multiple access (TDMA) will

be preferable to code-division multiple access (CDMA) in dealing with non-linearities.

4.4.3 Phase manipulation

Consider a phased-array antenna that emits two sinusoidal beams at the same frequency, but with different and known phase shifts. We take the beams to consist of a strong (label S) and a weak (label W) beam so that the total input signal is of the form

$$S_I(t) = A_S \sin(wt - \phi_S) + A_W \sin(wt - \phi_W) \tag{4-7}$$

where the phases are known. The weak beam could be, for example, the sidelobe of another beam in the main lobe of S.

Instead of going through all the complications of a Volterra model with memory, suppose that the output signal from an amplifier of nominal linear gain g has a term which is simply the cube of the input signal as in Eq. (4-1), so that the linear plus in-band term of $O(A_W)$ in IM distortion is

$$S_O(t) = A_W \sin(\mathrm{wt} - \phi_W)[g + \frac{3\epsilon A_S^2}{2}] + \frac{3\epsilon A_S^2 A_W}{4} \sin(\mathrm{wt} + 2\phi_S - \phi_W)] + \ldots. \tag{4-8}$$

Now one wants to get rid of the unwanted term with phase $2\phi_S - \phi_W$, leaving a remainder that reproduces the weak input signal except for the gain factor g. This can be done, even if the amplitudes ϵ and $A_{W,S}$ are unknown, by (digitally) multiplying S_O once by a local "oscillator" whose form is proportional to $\sin(\mathrm{wt} + 2\phi_S - \phi_W)$ and low-pass filtering, then multiplying S_O by $\sin(\mathrm{wt} - \phi_W)$ and low-pass filtering. Solving two simultaneous linear equations based on these results yields the necessary amplitudes of the signals with different phases in Eq. (4-8), so that the unwanted part with phase

$2\phi_S - \phi_W$ can be removed and the appropriate linear gain g applied.

Of course, this is a very simple example, but it can be generalized to more complicated situations; we do not discuss that here.

4.4.4 Nonlinear Undistortion

In the development of AWA one must maintain accurate linearity of the many elements so that the superposition of properly phased signals from many sources can be used to form beams of interest. The individual elements, however, may exhibit nonlinearities associated with the amplifiers, the solid state devices or other sources. Correcting these nonlinearities is often addressed using an idea of pre-distortion which shapes the input signal to the nonlinear devices with the expectation that these nonlinear signals will emerge properly linear as they depart the antenna.

Here we suggest another approach which uses signal processing at the receiver to adjust for the nonlinear mixing of CW signals from the transmitter. The idea is to use a set of "test signals" to probe the transmit, reflect, receive system in a calibration mode to determine a nonlinear state space filter whose weights can be set in the calibration phase. These weights are used in the transmit, reflect, receive mode and serve to "undistort" the signals as long as the system is stationary.

The formal setup is that we introduce a signal $s(t)$ into a device, a nonlinear device, which transmits an output signal $o(t) = h(s(t))$ where $h(s)$ is a nonlinear function characterizing the unknown nonlinearities of the transmitter. We, in effect, wish to invert the function $h(s)$.

The state space of a signal is determined by more than amplitude and

phase if the signal has a nonlinear source. The state of the system is determined by the other dynamical variables operating on any given input, DC or other, and from the measurement of a single output, one may reconstruct a "proxy" state space which characterizes the signal source. This is done routinely in nonlinear dynamics, and the ideas carryover here.

The basic operational idea is that one needs to sample the signal of interest, call it $s(t)$ as here, at several locations in time to determine how the many dimensional state of the signal source was projected onto the signal axis $s(t)$. One convenient way to do this is to form a proxy state vector $S(t)$ from the signal $s(t)$ and its time delays $s(t - nr)$ where n is an integer and τ is the sampling time at which we observe the signal-one over the sampling frequency. So one forms the proxy vector

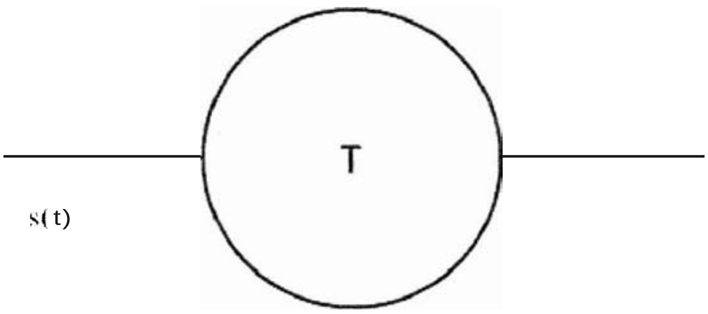
$$S(t) = [s(t), s(t - r), s(t - 2r), \cdots, s(t - (D - l)r)]$$

where rand D are determined in standard ways. Let us leave those determinations alone for the minute. The proxy vector S(t) has information in it about all the other variables in the nonlinear system because what happens to the signal τ earlier is part of the determining factors in the signal now. The D dimensions tell us how many earlier time delays we need to characterize the full state of the system now.

The idea is to construct the inverse of $h(s)$ <u>locally</u> in the appropriate proxy space rather than globally as is done in conventional predistortion schemes. This can succeed because in those proxy spaces the description of the state is unambiguous due to projection as we have essentially "unprojected" the state.

What do we do here? **In** the figure is our problem. $s(t)$ enters the transmitter and emerges not as $s(t)$ but as $o(t) \neq s(t)$. The state of the

signal source is characterized, as indicated above by the proxy vector $S(t)$, but it is also characterized by the proxy vector for the output O(t) calculated in just the same fashion using time delays of the output. Since either $O(t)$ or $S(t)$ each fully characterize the transmitter nonlinear operation, we can write $s(t) = g(O(t))$ where $g(S)$ is to be determined as we shall show in a moment, or we can write $o(t) = f(S(t))$. If, from $f(S)$, we now form $F(t) = [J(t), f(t - T), \cdots, f(t - (D - 1)T)]$ and similarly for $G(t)$, then $O(t) = F(S(t))$ and $S(t) = G(O(t))$, and they are inverse functions of each other in the D-dimensional space.



To identify the input to the transmitter which will give $s(t)$, the desired output, we should do this:

1. from $s(t)$ form the D-dimensional vector $S(t)$.
2. using the function $g(x)$, noting that 9 operates on D-dimensional vectors, form g(S). This is a scalar function of the proxy vector $S(t)$.

If we were to form $G(S)$, then after it passed through the transmitter it would produce $f(G(S)) = S$. This tells us that the signal we wish to present to the transmitter is $g(S(t))$, and for this we need to determine $g(x)$.

Go back to our original problem:, we present $s(t)$ to the transmitter and produce $o(t)$ as output. These are measured and recorded for a test

set of inputs. Form the D-vector $O(t)$, $g(O(t)) = s(t)$ which we also know. Now in the D-space, each $O(t)$ has neighbors, say N_B of them, and each of these maps with the same function $g(x)$ to nearby inputs, N_B of them. Call the neighbors of $O(t)(r)(t); r = 0, 1, \cdots, N_B; O(O)(t) = O(t)$ and $s(r)(t); r = 0, 1, \cdots NB; s(O)(t) = s(t)$, all of which satisfy $s(r)(t) = g(o(r)(t))$.

We can represent, at any time, the function $g(x)$ as

$$g(x) = \sum_{m=1}^{M} c_m(t)\phi_m(x),$$

where the functions are some set of basis functions, that could be polynomials, defined on vectors x in the proxy space. The coefficients can be determined by minimizing the least squares expression

$$\sum_{r=0}^{N_B} \left[O^r(t) - \sum_{m=1}^{M} c_m(t)\phi_m(S^{(r)}(t)) \right]^2$$

thus utilizing all the information about the neighbors too. This determines the function $g(x)$ in each past of the D-dimensional state space, and it does so <u>locally</u> in that space.

Now if we are given a new input signal $q(t')$ and we want to apply $g(x)$ to $Q(t')$ made in the usual time delay way, we need the $g(x)$ for the region of D-dimensional space where $g(x)$ operates. To locate this we look for the vector in the training set $O(t)$ closest to the vector $Q(t')$ and use the function $g(x)$ for the region of D-space near $O(t)$ to form

$$g(Q(t')) = \sum_{m-1}^{M} c_m(t)\phi_m(Q(t'))$$

as the input to the transmitter. This gives as output from the transmitter a value which is very close to $q(t')$ by construction.

Determining the number M of basis functions, the basis functions themselves, etc., is a matter of art and requires some experimentation.

While the calculations are numerous, one can calibrate each element of the array in this manner and preprocess the signal input to that element so that the output is the desired signal *s(t)* at time *t*. Combining these signals with proper delays to form the desired beam proceeds as usual.

This procedure has been used by the San Diego company, Chaos Telecomm., Inc., to determine the full nonlinearity in a DSL channel including impairments in the copper connecting the telephone office and the customers' premises and nonlinearities in the amplifiers at the telephone office. As that too is an OFDM problem, the methods may well apply to undistorting the AWA signals. Actually, Chaos Telecomm used a clever, still proprietary, implementation of the basic algorithms indicated here which allows real time processing with minimal computing power.

4.5 Findings - Digital Waveforming in Active Phased Arrays

The goals of multi-functional applications, rapid adaptation to evolving technology opportunities, and the ability to react readily to new threats are best supported by increasing the level of programmable (soft ware controlled) function in the system architecture. Rapidly developing commercial technologies appropriate to achieving these goals can be adapted in the Integrated Topsides Program with low technical risk.

- Commercial developments in RF-CMOS are changing the design rules that govern performance and cost for the front end electronics in low-power phased arrays. As a result the cost-drivers for multi-function

processing are changing rapidly with resulting opportunities for reduction in cost and increase in digital control.

- Frequency down-conversion of all the beams individually is possible because mixers are cheap in CMOS. This eases the constraints on ADC and DAC rates allowing digitization of the antenna signals before wave forming.
- Use of voltage-controlled oscillators, rather than expensive varactors, to create discrete beam phase shifts is efficient in CMOS. Subject to careful design to avoid addition of phase noise, this approach could serve as coarse phase shift in arrays with large number of elements, or could be cascaded through consecutive groupings of elements to perform the full phase shift function.

  Internal calibration of the relative phase shift among elements is possible within the front end electronics

  Economies of scale are necessary to realize the full cost advantage. Per-chip processing (CMOS) is inexpensive, but if custom designs are required, the mask sets presently cost ~\$500k, and mask cost will increase with decreasing length scale.
- SiGe BiCMOS has proven capabilities in production of low noise, wide-band electronics similar to those needed for low-to-moderate power active phased array applications.

• Given the potential cost/performance benefits of CMOS-RF front ends, an architecture with some combination of parallel transmit/receive processing elements for different **RF** functions at different frequencies, and tunable front-end frequency-band filters is likely to be the most feasible approach to wide band multifunction for low-power applications.

- The balance between the optimum level of analog and digital processing in the control electronics will continue to evolve over the next decade as materials and fabrication technologies advance. System design should remain flexible enough to support such evolution.

  - At present, full "software-controlled RF systems," e.g., individual synthesis and processing of the wave forms of all the elements at the control room is limited by the costs of high-rate data transmission and rates of *ADCjDAC* at high resolution.

  - The relatively low cost of digital signal processing does, however, allow designs with front-end digital signal processing for feedback, error correction, phase calibration and predistortion, and should substantively improve the performance that can be achieved for a given cost.

5 SYSTEM DESIGN

The two major cost drivers for active phased array systems are RF electronics, discussed in the previous 2 sections, and the overall system complexity. Adding to system design the requirements of reducing the number of apertures, relieving electromagnetic interference, and improving performance certainly increases the difficulty of the problem. In the following we address issues of system design for an Integrated Topsides program in the context of controlling the costs associated with system complexity.

5.1 Antenna Configuration

Configuring active phased-array apertures for multi-functional uses is a problem of minimizing aperture area within the constraints of supporting the diverse functional requirements illustrated in Section 2 Table 1. Two specific issues that must addressed in this problem are determining the level of "function-sharing" that is feasible for a given aperture, and the problem of configuring the antenna-elements within the apertures to support the wide band of frequencies (2-18 GHz). The potential use of analog beam steering (e.g., using hardware focusing) to minimize the complexity of the electronics system is also of possible interest.

How small the area allocated to active phased array apertures can be depends crucially on how their multi-function capabilities can be configured. Separation of the transmit and receive arrays is needed for signal isolation, and this approach was used in the AMRFC demonstration. Beyond this, the issues of how much multi-function can be handled on the same antenna

elements needs further technical analysis (see Section 5.1.1 for a preliminary analysis). As suggested earlier in this report, separation of high-power and low-power function is likely to be a good approach because of the different demands on the TRM amplifiers, signal dynamic range, and the possibility of interference or cross-talk. The approaches to multi-function apertures demonstrated in the AMRFC test-bed were quite different for transmit and receive arrays. The transmit array was divided into 4 quadrants of 256 antenna elements each. Four transmit waveforms up to 1 GHz in bandwidth were independently generated and could be routed to any or all of the four quadrants. **In** testing, multiple functions were not transmitted simultaneously from a single quadrant. Instead individual functions were switched between different quadrants, as illustrated in Figure 21. Switching was successfully demonstrated. Temporal switching allows some saving of aperture area, depending on the duty cycles of the different functions, and time multiplexing could yield further improvement where the functional requirements allow this. However, the greatest reduction in aperture area will occur when multiple functions can be transmitted simultaneously from the same element. From the successful integration of several receive functions discussed below, it seems quite likely that transmission of multiple low-power signals is practical. However, evaluation and experimental testing of the effects of combining different transmit functions on the same aperture elements are needed.

On receive, the AMRFC apertures are divided into 9 subarrays of 128 antenna elements each. These receive signals from these subarrays can be processed in combinations of three, six or all 9 subarrays. The receive modules are distinctly different than the transmit modules. On receive, the signal from each antenna element is split into 4 channels, each of which has a separate capability for phase-shifting and combination. The signals from the indi-

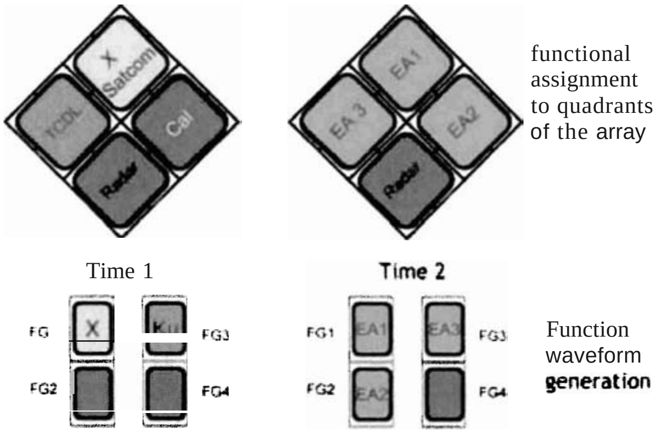


Figure 21: AMRFC transmit array configuration. The aperture is divided into four sub-arrays each of which can be temporally switched among different functions. Figure adapted from [4].

vidual sub-arrays can subsequently be routed to analog or digital (wideband or narrow band) processing circuits to complete the beam-forming process. In testing, analog combination was used for communications functions and narrow-band digital for navigational radar. Functional receive was demonstrated for four simultaneous functions, three communications in X and Ku band, plus low-power navigational radar in the X or Ku band. These results suggest that handling multiple functions simultaneously should be routinely achievable at low power levels. Issues of jamming or dynamic range, when dealing with combinations of weak and strong signals, could be mitigated by frequency selection for signals in different frequency bands.

The issue with configuring antenna elements within the array fundamentally is fundamentally limited by the requirement that inter-element spacing be no larger than $\frac{1}{2}$ of the wavelength for the highest frequency to be used, and that the width of the aperture (and thus the number of elements in the aperture) be large enough to focus beams to the desired narrow width. More detailed issues of the exact design of the antenna elements themselves and their geometrical distribution have higher-order effects on the array perfor-

mance [43]. In the AMRFC - LCS design concept, the elements are uniformly spaced on a square array, with apertures of different element spacing (2.75 cm for the 4-8 GHz range, and 1.0 cm for the 8 - 16 GHz range). Given the cost of the transmit/receive modules (one per function or frequency band, as discussed above, per antenna element), there is significant motivation for creating sparse array designs, as illustrated in Figure 23. Such designs can

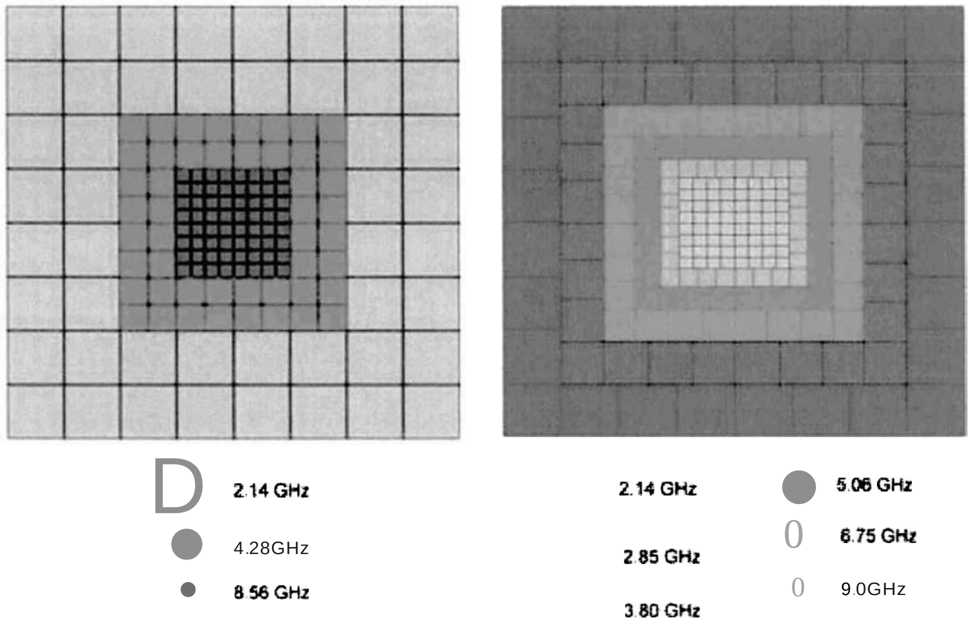


Figure 22: Schematic illustration of two sparse array concepts. Antenna elements would be positioned at the cross-points, allowing lower densities of antenna elements for the lower frequency bands. Figure adapted from [44]. Left: a 2:1 architecture, right: an architecture that allows a finer gradation of frequency bands while maintaining a constant number of elements at the edge of each subarray.

give excellent beam shapes for the different frequencies. However, hard-configuring a specific sparse/dense configuration should be avoided, as it will reduce the potential for flexibly reallocating array elements to different functions. We recommend that initial choices for array architectures maintain more flexibility for future evolution. For instance, the 2:1 architecture shown in the left of Figure 23 could be designed with open (unpopulated) antenna

positions at the highest density. The population of the array elements could, for instance, involve full poplulation with antenna elements of the same design, but differential population with TRM of different frequency bands, or it could involve antenna elements with different designs optimized for different frequency bands. More powerful designs are likely to be possible using optimization procedures. One example is the design of fragmented antenna arrays, for which optimization of aperture gain over a broad frequency band has been demonstrated [46]. In either case, hardware interface modularity (see Section 5.4) would be required to allow hardware reallocation of the antennas and/or TRMs, and thus relatively easy reconfiguration of the array.

5.1.1 Spurious beams from multi-function signals

A major question in evaluating efficiencies for multifunction arrays remains. This is whether a given set of antenna elements could be used simultaneously to generate multiple beams are transmit. In evaluating this possibility, a serious aspect of multi-function arrays will be the potential for interference arising among signals generated simultaneously on the same (or neighboring) antenna elements. As emphasized throughout this report, performance at the system level is the key performance criterion. Nonlinearities of individual components or modules can create harmful spurious effects by harmonic or intermodulation distortion ("spurs") as discussed in Section 4.4, but their impact on performance must ultimately be assessed at the system level. In this section we will study spurious beams from an entire array that may be created by spurs in the individual modules. It will turn out that some spurs do create propagating spurious main beams in the antenna pattern at the spurious frequency. In contrast other spurs do not, essentially because the would-be spurious beam is an evanescent wave, and only propagates in

some of its own sidelobes. This result should be useful in the design of advanced arrays in which several signals share a common array (or a common subarray). The importance for system design is that some spurs in individual modules must be stringently controlled, because they do create a spurious main beam, while we may be able to relax: the tolerance on other spurs, because they do not propagate as a spurious main beam. Thereby, system optimization may be easier. These considerations apply on both transmit and receive.

Consider a two-dimensional planar array of modular antennas, sufficiently closely spaced so as to Nyquist-sample the waves under consideration ($k_T d < \pi$, where d is array spacing and k_T is the magnitude of the component of wavenumber transverse to the plane of the array). Use cartesian coordinates (x, y, z) where the array lies in the xy-plane. The array will be simultaneously excited by one or more main beams at frequencies ω_i $(i = 1, 2, 3 \dots)$ and corresponding transverse wavenumbers k_i; these signals are assumed CW. By nonlinear mixing, a spurious signal will appear at some harmonic or beat frequency ω; for instance we might have spurs at $\omega = 2\omega_1$ (second harmonic), or at $\omega = \omega_1 + \omega_2 - \omega_3$ (third order intermodulation). At the same time the spur will have a transverse wavenumber k_T, which would be $k_T = 2k_{T1}$ (second harmonic) or $k_T = k_{T1} + k_{T2} - k_{T3}$ (third order intermod), for example.

Under what condition will the spur ω propagate as a main beam? Let k_z be the component of wavenumber perpendicular to the plane of the array; so $k_z = \mathbf{k} \cdot \hat{\mathbf{z}}$ where $\mathbf{k}$ is 3-dimensional vector wavenumber of the main beam, and $\hat{\mathbf{z}}$ is the unit vector in the z-direction perpendicular to the array. Then by the dispersion relation for a free wave, $(\omega/c)^2 = |\mathbf{k}|^2 = |\mathbf{k}_T|^2 + k_z^2$, we

obtain

$$k_z = \sqrt{(\omega/c)^2 - \mathbf{k}_T^2} \tag{5-1}$$

for the wavenumber in the z-direction, and the main beam will propagate only if k_z is a real number, *i.e.*, if $k_T < \omega/c$; here $k_T = |\mathbf{k}_T|$ denotes the magnitude of the component of wavenumber k transverse to the plane of the array. In the opposite case, the main beam will be "evanescent" (to borrow a pleasing term from optics), that is, it will die off exponentially like $\exp\left(-z\sqrt{|\mathbf{k}_T|^2 - (\omega/c)^2}\right)$ in the z-direction perpendicular to the array. When evanescent, the main beam represents reactive near-field energy, not a propagating beam.

However the array will have sidelobes (as can be computed by the usual linear theory), and at the spurious frequency ω, some of the sidelobes will propagate rather than being evanescent. For instance, in a simple I-dimensional array without taper, the propagating spurious sidelobes would be lower in amplitude by a factor

$$\frac{2\omega}{ck_T + \omega} \frac{\sin((ck_T - \omega)L/2)}{(ck_T - \omega)L/2} \tag{5-2}$$

or smaller, compared to the evanescent spurious main beam, when $k_T > \omega/c$. In a detailed analysis, effects of the spurious sidelobes would also have to be considered.

Now we turn to the properties of the actual spurs, and to the question of whether the corresponding spurious main beams are propagating or evanescent. Nonlinearities in the modules can be expanded as a power series in total amplitude, to separate out order-n effects, where $n = 1$ means linear, $n = 2$ means quadratic, and so on. The nonlinearities may have memory, which will simply change the phase of the spurious nonlinear signals under our assumption of pure CW drive.

In order $n = 2$ we can get second harmonics $\omega = 2\omega_1$ of a single input ω_1; we can also get sum and different beat frequencies $\omega = \omega_1 \pm \omega_2$. The former is just a special case of the latter, with $\omega_1 = \omega_2$. The total system bandwith must span at least an octave for the spurs to'appear in-band. The corresponding wavenumbers are $k_T = k_{T1} \pm k_{T2}$. Using the dispersion relations, we derive for the spurious main beam

$$k_z^2 = (k_{1z} \pm k_{2z})^2 \pm 2k_{1z}k_{2z}\frac{1 - \mathbf{n_1} \cdot \mathbf{n_2}}{\cos\theta_1 \cos\theta_2} \tag{5-3}$$

Here the first exciting beam ω_1 is propagating in the direction of the unit vector $\mathbf{n}_1$, at an angle θ_1 to the z-axis, with $\cos\theta_1 = \mathbf{n}_1 \cdot \hat{\mathbf{z}}$, so $k_{1z} = |\mathbf{k}_1| \cos\theta_1 = \omega \cos\theta_1/c$; and similarly for the second exciting beam. From the discussion above, the spurious main beam at ω will propagate if the right-hand-side is positive, otherwise not. It follows that, for *sums* of exciting frequencies $(\omega_1 + \omega_2)$, and for second harmonics, the spurious main beam *always propagates,* because the right-hand-side is the sum of positive terms. On the other hand, for difference frequencies $(\omega_1 - \omega_2)$, the main spurious beam is often nonpropagating, depending on frequencies and angles; detailed evaluation is required to decide. As one example, if the two exciting beams are each 45° off boresight, and 90° from one another, then

$$k_z^2 = (k_{1z} - k_{2z})^2 - 4k_{1z}k_{2z} = k_{1z}^2 + k_{2z} - 6k_{1z}k_{2z} \tag{5-4}$$

and the main beam will be evanescent as long as $\omega_1 < (3 + 2\sqrt{2})\omega_2 \approx 5.8\omega_2$.

Order $n = 3$ may be treated in a similar manner though the details become more complicated. The most interesting case is the combination

$$\omega = \omega_1 - \omega_2 + \omega_3 \tag{5-5}$$

which is allowable even in narrow-bandwidth systems. A special case is the combination $\omega = 2\omega_1 - \omega_2$ of two excitations. The transverse wavenumbers

combine as $\mathbf{k}_T = \mathbf{k}_{T1} - \mathbf{k}_{T2} + \mathbf{k}_{T3}$ and the dispersion relations give for the wavenumber in the z-direction

$$k_z^2 = (k_{1z} - k_{2z} + k_{3z})^2 \quad -2k_{1z}k_{2z}\frac{1-\mathbf{n_1}\cdot\mathbf{n_2}}{\cos\theta_1\cos\theta_2} + 2k_{1z}k_{3z}\frac{1-\mathbf{n_1}\cdot\mathbf{n_3}}{\cos\theta_1\cos\theta_3} \qquad (5\text{-}6)$$
$$-2k_{2z}k_{3z}\frac{1-\mathbf{n_2}\cdot\mathbf{n_3}}{\cos\theta_2\cos\theta_3}$$

We have not attempted to determine general conditions for the existence of propagating spurious main beams; this would be a lengthy but straightforward analysis. Two parametrized examples will be enough to show that both cases do appear. Take the three exciting beams all nearly of the same frequency, $(\omega_1 \approx \omega_2 \approx \omega_3)$ and all at the same angle from boresight, $\theta_1 = \theta_2 = \theta_3$, though in generally different directions $\mathbf{n}_i$. Then this expression reduces to

$$\frac{k_z^2c^2}{\omega^2} \approx \cos^2\theta_1 + 2(\mathbf{n_1}\cdot\mathbf{n_2} + \mathbf{n_2}\cdot\mathbf{n_3} - \mathbf{n_1}\cdot\mathbf{n_3} - \mathbf{1}) \qquad (5\text{-}7)$$

If we arrange the excitations so that beam 1 is 90° from beam 2, and beam 3 is also 90° from beam 2, but 1 and 3 are less than 120° from each other, then the right-hand-side is clearly negative, and the spurious signal ω is evanescent. On the other hand, if 1 and 3 are each less than 60° from 2, but are 90° from each other, then the right-hand-side is positive, and the spur will propagate as a main beam. Thus for third-order nonlinearities, both cases occur, and in fact both are widespread. More detailed analysis would be necessary to go much further.

The overall point is that nonlinear distortions need not be completely eliminated; it suffices to put their energy out-of-band, *or* to put their energy into modes which do not propagate out of the array, i.*e.*, into evanescent array modes. The type of analysis presented here can be used to evaluate the issues of non-linear distortion in system designs involving multiple simultaneous function supported on the same array elements. It also can be used to develop

operations criteria (e.g., relative directionality of simultaneous beams) to minimize effects of non-linearity. Another point is that although evanescent waves decay exponentially and do not propagate to large distances, they can couple to receivers and other electronics at short distances, e.g., on the same ship, especially if the source is at a high power level. Thus, evanescent modes may not be completely benign and require assessment, especially for high power Tx functions.

5.1.2 Analog beam steering

Beam forming historically developed as a natural evolution of the hardware used in rotating dish antennas. That is, capture of the full angular spread of the incident radiation, with subsequent electronic processing to extract the angular information. There is an alternative to this approach based on the traditional methods of optics and diffraction. This is to use a lens system to physically separate radiation incident at different angles into different physical locations. Specifically, the well-known result of paraxial ray optics is that all radiation incident on a lens at a given angle is focused to a single point in the focal plane. This property can also be used at radio frequencies, subject to the issue of aberrations, which are universally present in such focusing systems.

Radio frequency radiation can be focused by using the refractive index (or impedance) properties of materials physically shaped [45, 47] in way similar to traditional glass optics. In addition, effective lensing can be accomplished by using spatially-configured time delays [48, 49]. The latter method is accomplished by using a planar array of antennas coupled through hard-wired time-delay links to a complimentary planar set of antennas, as

illustrated in Figure 23. One plane receives a signal, generates an electrical signal, which is sent to the second plane through the time delay, and then the second plane of antenna elements transmits the signal. On transmit, the feed source is placed at a focal position chosen to matched the desired angle for the outgoing beam. On receive, beams incident at different angles are collected at different focal positions.

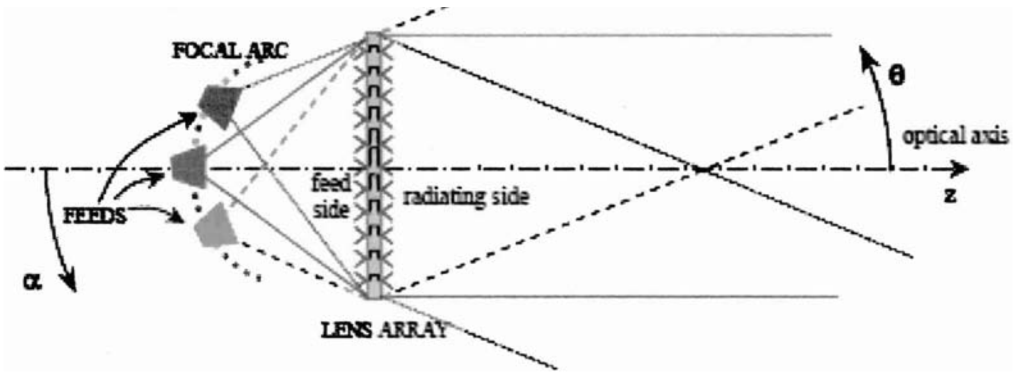


Figure 23: Schematic illustration of a discrete lens array operated in transmit mode. The feeds are located on the focal arc, the position of which is determined by the configuration of matched time delays between the input/out planes of antenna elements in the lens array.

Analog beam-forming has the benefit that electronic processing of the incoming signals from all the different antennas is greatly reduced (e.g., adding phase shifts and combining is no longer needed.) Instead, the lens effectively imposes a fixed set of phase shifts, and the fully formed beam is collected directly. The drawbacks of such a configuration are two-fold. First, one looses the possibility of controlling the individual phase information, with all the flexibility, automated calibration/correction, and powerful processing that includes. Second, one is subject to the physical limitations of aberrations in focusing. **In** the case of the discrete lens array, careful tailoring of the time lags between elements allows aberration correction for handling large areas. However, the specific designs are wavelength-dependent, which means that

the lens will be subject to chromatic aberrations if frequencies much different than its design specification are used.

Analog beam forming is certainly feasible, although at a cost of substantially more complex aperture design. That drawback may be compensated by the potential for reducing the overall complexity of the beam steering system. There is also another proposed antenna configuration, reconfigurable apertures [?], that has the potential for significant reduction of system complexity, while still maintaining the benefits of the fully digital interface. The latter approach is also supported by substantial commercial interest in reconfigurable apertures for improved channel capacity [51], and thus developments in this area are likely to be rapid. We would suggest a thorough evaluation of the suitability of both analog beam forming and reconfigurable apertures for Integrated Topside applications, based on the following criteria in comparison with the more traditional configuration:

1) Lens array volume and profile

2) Quality of beam shape for a single lens design over a wide range of frequencies

3) Improvements, if any, in sensitivity to amplifier linearity

4) Cost gain due to reduced electronics complexity

5) Robustness of lens array configuration under field conditions.

5.2 Calibration and Validation

The need for calibration/validation of phased array antennas is well known, e.g., for communication and radar satellites [52, 53]. For an agile,

multifunction, wideband aperture this need is more pronounced because of the multiple functions that the array is to perform, e.g., communication, radar and electronic warfare. The calibration must satisfy the most demanding requirements of the several functions of the aperture. Thus, the calibration/validation function of the aperture should be able to assure performance for pointing, gain and beam shape as well as limiting unwanted emissions from sidelobes and nulling interfering signals for receive functions. Calibration is important from another aspect if one uses delta-sigma schemes (see Section 4.2.1) for conversion between digital and analog versions of a signal. Probably the most important aspect of calibration is to discard data from failed elements and compensate for the missing data in the beam forming and other aperture functions. Editing out bad data and compensating for it in signal processing allows one to mitigate the impact of failed elements.

Calibration takes place at the subsystem level and more importantly at the system level as end-to-end testing and optimization of aperture performance. Subsystem level calibration is valuable even if end-to-end calibration is also done because the former can be done more frequently and simplifies the task of optimizing performance using data from end-to-end testing. Probably the most important calibration is for the phase reference signal that goes to each element. For a wideband aperture this is typically an analog signal and thus subject to distortion by multiple reflections in the signal pathway. We note that such a reference signal can be auto-calibrated eliminating the need for any delay matching by connecting phase signals from adjacent elements together. During calibration, each element would measure the relative phase between it and its four neighbors and calculate a phase correction to bring them in phase. After a few iterations of this phase correction all modules would converge on a common "zero phase" without the need for any

matched-length cabling of the reference signal. Subsystem level calibration that is both integrated and diverse (20 modes) has been implemented in the AMRFC Test Bed [4].

End-to-end calibration and validation is a critical element in the design of multifunction apertures since many functions depend on only a few apertures. We will discuss two approaches to end-to-end testing of multifunction apertures by transponder/beacons and special sensing elements in the transmit and receive apertures. End-to-end calibration is probably best performed by external transponders (for transmit and receive arrays) and beacons (for receive arrays). The concept is illustrated in Figure 24 below. A transponder or beacon could be on board the same ship or off board on a buoy, aircraft, UAV, spacecraft or other ship. The need for such external calibration/validation of aperture performance is to make sure that effects beyond the last stage of subsystem calibration are taken into account, e.g. changes in antenna elements (notch structures) or interconnects by weather or other damage, propagation, scattering from other structures on the ship, etc. The objective is to assess aperture performance and compensate for failure, degradation or other changes in aperture components. Some of the functions that can be assessed and possibly improved are as follows:

- Beam pointing
- Beam shape
- Sidelobe levels
- Unanticipated sidelobes
- Gain.

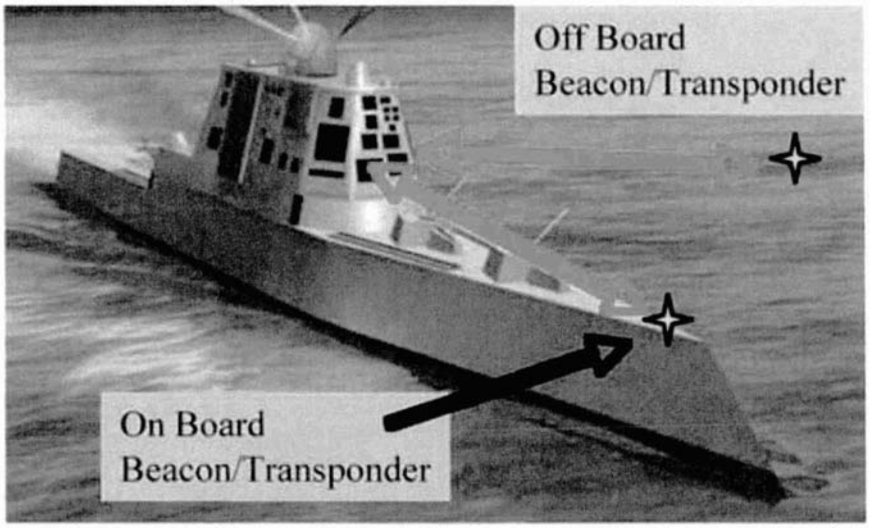


Figure 24: Calibration/validation of aperture performance by end-to-end testing of aperture function. Off board and on board transponders and beacons are illustrated.

There is a wide variety of measurements that can be made with transponders and beacons to assess and optimize phased array antennas. Spacecraft phased arrays can be assessed and optimized using transponders or beacons on the ground [52]. Using a transponder or beacon one can scan antenna beams across the point source to observe the antenna pattern and optimize it. Clearly one would like to probe antenna beams in all directions since the calibration could change depending on the direction that the aperture beam is pointing. On board transponders or beacons are easier to implement, but are limited to a set of directions that generally doesn't include the directions that the aperture is most concerned with. Nevertheless, on board transponders/beacons can indicate failed elements and provide a useful assessment of aperture performance. Fenn [54] argues that on board transponders do not have to be located in the far field of the aperture as near-field measurements can be used to infer far-field measurements by exploiting the near-field focusing capability of a phased array aperture.

Off board transponders/beacons are more flexible, but harder to implement. Use of a UAV with a transponder and beacon appears to offer a cost-effective way to make maximum use of off board calibration. Several advantages of off board calibration/validation are as follows:

- Can be done in many diverse directions, including high elevation angles
- Can easily be done in the far field of the aperture
- Includes more interaction of aperture with ship structures.

The biggest disadvantage of off board calibration is the difficulty of implementing it while underway or in bad weather. The required apparatus and platform for off board systems are more expensive, but a small part of the overall cost of a multifunction aperture.

Another approach to end-to-end calibration/validation is to use the nearby elements in the aperture array as sources and/or sensors. In the case of radar arrays Auman et al. [55] and Shipley and Woods [56] have discussed using the mutual coupling between transmit/receive (T/R) modules. This technique is sometimes called, mutual coupling auto-calibration' or just 'auto-calibration.' The former authors report an experiment in which the method was applied to a 1.3 GHz array with rather good results. Beam shape was within 1 or 2 dB for the antenna pattern within 30 dB of the peak and within about 5 dB down to 40 dB below the peak. It appears that future multifunction apertures for integrated topside are likely to have separate transmit and receive apertures. Thus, one would need to put some transmit elements in the receive aperture and some receive elements in the transmit aperture as shown in Figure 25 below. A small-scale version of this

scheme was implemented on the AMRFC test bed, but has not been used for calibration thus far.

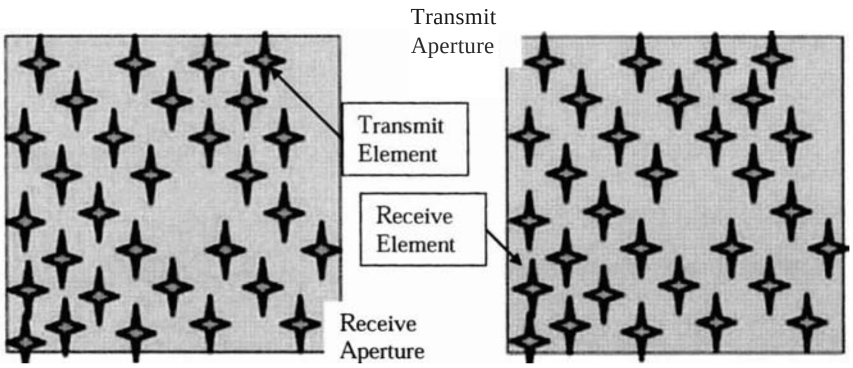


Figure 25: Sampling of aperture performance with transmit elements (red stars) distributed over the receive array (blue) and with receive elements (blue stars) distributed over the transmit aperture. This scheme was implemented on the AMRFC test bed, but not used for calibration.

The auto calibration method is similar to the near-field antenna measurement method that has become popular as it is more economical than far-field antenna ranges, (see Yaghijian [57]). Rahmat-Samii [58] describes a near-field test facility at UCLA. One could extend the mutual coupling method by introducing probes that would perform near-field antenna measurements better than simply using extra elements in the aperture. However, such an extension of the mutual coupling method would require significant added expense in both the apparatus and the time required to make such measurements. The auto calibration method is less useful in assessing overall system performance than an airborne transponder/beacon would be. However, it has the advantage of being relatively simple to implement and would be available for frequent use.

Whatever means one uses to collect data for aperture performance, these data must be processed to make a useful assessment of aperture performance

and enable performance improvement if needed. As a simple example, suppose that a particular element has failed and is supplying anomalous data. The data analysis algorithm should recognize the failure and identify the failed element. Once this has occurred the faulty data can be ignored and steps taken to compensate for the loss of data from this particular element or elements. A straightforward way to compensate for the loss of a given element is to change the weighting of other elements to produce as good an antenna beam as possible by giving the faulty element zero weight and changing the other element weightings in the beam formation algorithm appropriately.

Approaches for making use of subsystem, element-to-element, beacon and transponder data have been developed with the twin objectives of assessment and performance improvement through remediation of faults and adjustments after calibration. One approach has the end-to-end philosophy and assesses the performance of phased array antennas that attempt to null interfering signals by placing a beam pattern null in the direction of the interference source. A near-field beacon that simulates a noise source can be used to estimate far-field nulling performance using near-field focusing of the phased array. By clever calibration procedures one can economize on the amount of calculation it takes to translate data into assessment and identification of faults and maladjustments. For example, Sorace [52] has developed a scheme for calibration of a phased array while it is in service with a minimum number of measurements. He argues that calibration at four orthogonal phase settings can be used in a maximum likelihood method to estimate accurately the calibration offset. Existing developments were not focused on multifunction apertures and more cost-effective methods that deal with the new aspects of multifunction apertures may be required after assessment of existing methods.

In summary, the need for continual calibration/validation in phased array antennas is well known and becomes more pronounced for multifunction apertures since the different functions have different requirements. The need for assessment of aperture performance focuses on the user who needs continual assurance that such a complex system as a multifunction phased array is working properly. Seeing a rotating dish antenna through the window assures a commander that the system is likely working. Such real time performance assessment is needed for multifunction apertures because of their importance and users unfamiliarity with them. This is especially so in the transition period when multifunction apertures are coming into use. We suggest a calibration approach which has the following steps moving from less effective, but easily implementable methods, that are used very frequently to more effective methods that are harder to implement and thus used less frequently. A comprehensive program could work as follows:

- Very frequent calibration/validation of subsystems and end-to-end testing using element-to-element coupling
- Less frequent calibration/validation using transponders/beacons on board the ship
- Occasional calibration/validation using off board transponders/beacons mounted on DAVs or other platforms.

5.3 Scheduling

5.3.1 Introduction

Managing the plethora of **RF** functions on a Navy vessel clearly needs careful advanced planning [1, 59,60]. A key feature of the multi-function concept is the use of software-controlled management tools to allocate shipboard **RF** resources optimally and dynamically among the functional requirements. The AMRFC program [59, 60, 4] has demonstrated the feasibility of dynamic function management in the Chesapeake Bay testbed, and has pioneered a modular open architecture suitable for reconfiguring the system in real time, and to adapting the system to future changes as capabilities and functional needs evolve.

The scheduling issues involve prioritization among the **RF** functions, given the resources available. Such prioritization will generally begin with a most common standard priority order, for which specific priorities would be adjusted in real-time to adapt to predictable variations in situations (e.g., ship in port, vs. convoy, etc.). Insuring that each function is accomplished requires experienced knowledge of the situations likely to be encountered and the corresponding **RF** demands generated by those situations (e.g., a rather static set of **RF** requirements when refueling at the home base, vs. rapidly changing **RF** needs when involved in a military exercise or combat).

To meet the time-varying functional requirements, the size, capabilities and physical distribution of the phased array apertures must be carefully planned. There will be some number N of apertures, with spatial locations that have different applicability for different functions. Each of the apertures

will contain an internal spatial distribution of M_N antennas, combinations of which will be software selectable for use in generating a signal. The choice of spatial combinations will be constrained by the frequency and beam width required for the function, and by possibilities of EMI or side-lobe formation due signals allocated to nearby antennas. Choices of functional allocation may be further bounded if individual antennas are (depending on cost limitations) differently equipped in terms of the functionality of their transmit/receive modules. Specifically, TRM choices will include the maximum power level, whether multiple beams of the same frequency will be independently steered using common elements, and whether multiple beams of different frequencies will be generated simultaneously using common elements.

Given the rapid evolution expected in the hardware capabilities for multi-function phased arrays, it is clearly not feasible (and certainly would not be wise) to attempt to generate a fixed solution to scheduling management. On the other hand, constantly evaluating the scheduling possibilities *as* hardware solutions are being evaluated is essential to optimizing the overall system design. Full cost-benefit analyses can only be accomplished in the context of how resource allocation can be optimized.

In the following, we describe in a general sense the issues involved in developing scheduling solutions. The importance of quantifying the physical and functional boundary conditions cannot be over-emphasized in such an analysis. In addition, we emphasize the importance of pre-defining resource allocations for anticipated scenarios, minimizing the need for real-time computational analysis.

5.3.2 Optimal Resource Allocation

As noted above, multi-functional scheduling requires allocating antenna elements from the arrays to various functions. Each of these functions has its own requirements (boundary conditions), including its temporal duty cycle, the frequency, the spatial positions of the apertures, and the number, spacing and TRM capabilities of the antenna elements required..

A resource allocation problem [61] is one in which we have resources $x_1, \cdots, x_n$ and wish to minimize (or maximize) some function $f(x_1, \ldots, x_n)$ of those resources subject to

$$\sum_{i=1}^{n} x_i = N$$

and

$$x_i \geq 0, i = 1, \ldots, n.$$

A discrete resource allocation problem is one in which the resources are integer valued, for example an integral number of antenna elements must be used. In general, it is computationally intractable to find optimal solutions to these problems except for small instances.

The discrete resource allocation is known to be NP-hard (NP stands for *non-deterministic polynomial-time)*. An NP-hard problem is a decision problem that is a member of a class H where every problem in the class NP is polynomially reducible to a problem in H. More formally, H is the class such that for every decision problem $L \in$ NP there exists a polynomial-time many-to-one reduction to H, written $L \leq_p H$. A language L is NP-hard if $\forall L' \in$ NP, $L' \leq_p L$. If a decision problem D is NP-hard and $D \in$ NP then D is NP-complete.

A problem is shown to be NP-hard by finding a polynomial-time (or logarithmic space) reduction to a problem that has already been shown to be NP-hard. In the case of the discrete resource allocation problem the reduction is to the *set partitioning* problem [61, 62, 63].

The set partitioning problem is given an m x n binary matrix $A = \{a_{ij}\}$ and a positive integer N, determine whether a vector $\vec{x} = (x_1, \ldots, x_n)$ satisfies

$$\sum_{j=1}^{n} x_j = N,$$

and

$$\sum_{j=1}^{n} a_{ij} x_j = 1, i = 1, \ldots, m$$

and x_j are non-negative integers, $j = 1, \ldots, n$.

It is strongly believed that there are no deterministic polynomial-time algorithms for problems that are NP-complete. As a result, the best algorithm that can be devised that provides an exact solution must be $O(2^n)$ (requires exponential time in the size of the problem). The proof of this assertion would be one of the most significant results in complexity theory, but several decades of effort provide strong evidence that it is true.

The problem of optimally assigning antenna elements to RF functions at appropriate time slots is similar to a discrete resource allocation problem called the *classroom assignment problem* [64]. In this problem there are a set of courses that to be taught (transmit/receive functions) each with a required enrollment of a certain number of students who have specific prerequisites (number and capabilities of antenna elements required for a function) and a set of classrooms each with a given number of desks (apertures with antenna elements, with the additional complication that individual antenna elements, or desks, may be used simultaneously for different functions, or classes subject to geometric constraints). There is an additional degree of freedom in the

time that the courses can be scheduled in order to find an appropriate classroom (including appropriate desks). The function to be maximized might be the number of courses that can be taught, subject to stringent priorities on the courses, or it could be the efficiency of utilizing the classrooms (analogous to power needed or EMI).

It is clear that finding an optimal solution to the multi-function phased array scheduling problem in real-time is not practical since it is NP-hard and for a non-trivial number of tasks the run-time will be prohibitive. This has already been recognized in the AMRFC software development program [60], where the allocation problem has also been bounded by pre-definition of antenna-element subsets that are allocated sequentially in time as dedicated groups to the prioritized functions.

In accomplishing maximization of resources use, there are important uses for finding optimal or near-optimal task schedules. The first is to find a set of fixed schedules in an off-line fashion that can be used as appropriate to the mission at hand. These schedules would be computed in such a way to provide the most efficient use of resources and provide the maximum number of free antenna elements for the longest possible intervals. These elements could then be used for best effort tasks such as communication, while tasks with fixed communication requirements would be guaranteed to have the resources that they need. The second use of finding optimal schedules could be as a comparison for real-time scheduling algorithms that might be developed to provide dynamic scheduling of the tasks. In general, dynamic schedules that guarantee that all tasks have the resources necessary to complete and can complete by the time required will be less than optimal. There needs to be some method of judging the efficacy of these algorithms, and comparing them *a posteriori* with known optimal schedules.

5.3.3 Real-time scheduling

Allocation of resources to multi-function phased arrays is an instance of a real-time scheduling problem. It is composed of hard real-time, soft-real time, and best effort tasks that must be completed in a timely fashion on a limited set of resources.

A task is said to be a *hard real-time* task [65] if a completion after its deadline can cause catastrophic consequences to the system. A task is said to be a *soft real-time* task if missing its deadline decreases the performance of the system but does not affect its correctness. A *best effort* task is one that has no deadline and can complete at any time. In general, the system should be starvation-free, that is, these tasks should be completed in a timely fashion.

A example of a hard real-time task in the AWA is tracking and jamming an incoming missile. An example of a soft real-time task is audio or video communication, perhaps via a TCP/IP connection.

Audio and video are soft real-time time since the quality of the stream (as measured by its jitter) is important to understanding the message, but a small amount of jitter does not cause a complete failure. Some examples of best effort tasks are sending a short message or an TCP/IP data connection.

In general, a scheduling problem can be defined as three sets: a set of n tasks $\{J_1, \ldots, J_n\}$, a set of m processors $\{P_1, \ldots, P_m\}$ and a set of s resources $\{R_1, \ldots, R_s\}$ [65]. In the case of the AWA problem, the tasks could correspond to transmit/receive events, the processors to antenna arrays, and the resources to antenna elements and other resources necessary to complete the task.

Real-time tasks, like other tasks, may not be executable in arbitrary orders but subject to precedence constraints. These precedence constraints can be described as a directed graph G with vertices J_i that represent tasks and directed edges that describe the ordering constraints among the tasks. $J_a \prec J_b$ means that J_a is a predecessor of J_b and the precedence graph G contains a directed path from J_a to J_b. $J_a \rightarrow J_b$ means that J_a is an immediate predecessor of J_b and the precedence graphs G contains a directed edge from J_a to J_b.

The graph G imposes a partial ordering on the execution of the tasks, and so constrains the scheduling algorithm. A task J_a with the constraint $J_a \prec J_b$ means that J_a must be executed before J_b, while if $J_c \not\prec J_d$ then J_c and J_d can be scheduled concurrently. The result is a significant increase in complexity of the scheduling algorithm, and simple greedy algorithms are unlikely to be sufficient.

There is a rich literature on real-time scheduling including texts [66, 67, 68] as well as many research articles. We recommend that this literature be carefully explored, as it is directly applicable to the AWA problem. Recent work on integrating hard real-time, soft real-time and best-effort scheduling [69] holds great promise to simplify the implementation of mixed scheduling systems, and is highly relevant to the AWA problem.

5.3.4 Single point failure

The AWA system is a complex one made up of many components, both hardware and software. It is essential for the success of the mission as well as the safety of the crew that the system function reliably and be protected against single-point failure modes, which might arise either due to faults in

the system or to damage occurred due to military operations. The reliability of the system can be enhanced through the use of best engineering practices, including software engineering practices, and the use of quality components. But it is important to remember that all hardware components can fail, and that *all* complex software systems have errors. In terms of software, it is simply impossible to construct a complex software system that is error-free, and any claims that such a system can be built should be regarded with extreme skepticism. Finally, under conditions of rough weather or combat operations, disabling damage may occur.

To avoid single points of failure in the system it is essential to make effective use of redundancy, and where appropriate, diversity to avoid common failure modes. It is essential that in the event that any component fails, the system performance degrade in a graceful manner. There should be no point in the system where the failure of a hardware or software component causes the entire system to fail. If the system is thought of as a directed graph, this means that there are redundant paths through the graph so that the loss of any single edge does not partition the graph. In terms of hardware, this means that there should be redundant power and communication channels. In terms of software, it means that best practices are followed and exceptional conditions are caught and handled rather than allowing the system to fail.

Fault-tolerance depends on the interaction of multiple components when a single one fails. The large number of components results in complex failure modes that may not be evident simply by examining the architecture. By using a combination of analytic modeling and simulation a better understanding of the complex failure modes of these systems can be obtained.

In the case of simple dependencies, a probabilistic analysis based on combinatorial models can be done. This analysis can be often be used to analyze simple subsystems and the result used to parameterize models of larger portions of the system. In the case of more complex interactions, it may be more appropriate to use analysis based on Markov networks [70]. The form of these networks are completely dependent on the interdependencies of the components, but the field is well developed and has been applied to many fields including complex computer networks.

When the system is so complex that an analytic solution is not possible, then discrete event simulation [71] can be applied to understand the interactions of the various parts of the system and estimate the reliability of the entire system.

5.4 ModularityjOpen Architecture

5.4.1 Introduction

The AWA system is a complex architecture that will be made up of components from many vendors and must be able to evolve over time as new technologies are developed and new capabilities are deployed. By designing the system as an open systems architecture, the Navy will not be captive to a single vendor and can take full advantage of the advances that are made in the commercial sector.

Even though the AWA system will be an open systems architecture, it remains an extremely complex system in terms of both hardware and software. In order for the pieces of the system to function in an integrated

fashion, in contrast to a collection of separate systems with little or no integration, the architecture must be well-defined, adaptable, and have standard interfaces and APIs (Application Program Interfaces) that can accommodate components from multiple vendors and evolve over time.

Examples of complex architectures exist in the commercial sector, and are increasingly common. Any complex system including the Internet, automobiles, and aircraft is composed of many components from many vendors that fit together in an integrated fashion. Aircraft, for example, are highly integrated in their electronics and avionics. A modern aircraft is assembled from components from many vendors around the world, and these components must fit together seamlessly.

As described in other sections of this report, to accomplish cost savings and most rapid implementation of new capabilities, we strongly recommend that wherever possible, the hardware used in the AWA system be standard COTS and will already have standardized interfaces. These interfaces were defined through standards organizations such as the IEEE, and are the reason that modern electronics such as computers can be assembled easily from components and require no additional circuitry. Adding a video card or a network interface to a computer is no more difficult than inserting it into a slot and installing the appropriate driver software. To the extent that it is possible, the multi-function phased array systems should be based on standardized commercial interfaces. Where necessary custom hardware standards specific to the system should be developed. In the software realm, there are many examples of open architectures that allow the system to interoperate and be extended. Notable examples include the the TCP/IP protocol suite from the networking community, the open source community, and even the proprietary Microsoft operating system (the driver architecture, for exam-

ple, allows for new devices to be added to the system without modifying the underlying architecture). As with hardware, processes exist for developing these standards based on consensus.

In the case of the AWA system, there will be a significant fraction of the hardware and software that is designed specifically for the system. This will be due to performance requirements or due to the uniqueness of the mission. The use of custom components does not mean that the interfaces should be unique to that component and to the vendor that produced it. A standards process similar to those used in the commercial world should be employed.

The standards processes that have worked well in practice are based on discussion and consensus. Having the Navy specify the architecture and all of the interfaces will not work well, since the Navy has limited technical manpower and the task would most likely be assigned to a contractor. The result would be an architecture designed by that contractor with the interfaces imposed by fiat. The Navy should focus its efforts on the development of requirements, which will include capabilities but should go beyond that and specify that the architecture be open and extensible. Interested parties, including contractors, can then be involved in a process that defines the architecture and develops standards for all of the hardware and software interfaces. In the following we present two examples of successful management of open architectures and hardware standards.

5.4.2 Development of Open Architecture

The software developed for the AMRFC testbed ([72], Section 5.3.1) provides an excellent example of an extensible open architecture, and should be used as a key component in developing the formal protocols for future

integrated topsides development. Furthermore, we note that the Navy has developed excellent open architecture standards [73, 74] that will further serve as a basis for the program.

We consider the Internet Engineering Task Force (IETF) an excellent example of the type of mechanism that the Navy should use in defining the Open Architecture standards for multi-functional phased array RF systems. The IETF is the group that develops and promotes Internet standards. It is responsible for the ubiquitous TCP/IP protocol suite, and for all of the protocols that make up the Internet. It is organized as a large collection of working groups, each of which deals with a specified topic. These groups are created to deal with a specific topic and once that topic has been addressed the working group disbands. Each working group has a chair and a charter that describes its tasking. The working groups are organized into subject areas and these areas are governed by an area director who appoints the working group chairs. The area directors along with the IETF chair for the Internet Engineering Steering Group (IESG) which is responsible for governance of the IETF. The IETF working groups operate fairly informally, reaching consensus through discussions and mailing lists [75] and has a set of policies for dealing with intellectual property rights [72].

The software system for the AWA architecture might be addressed in a similar way. By developing well thought out, and where appropriate, physically based interfaces the AWA architecture will be flexible and able to evolve as new capabilities become available.

5.4.3 Hardware Standards

The development of custom standards for the multifunction phased ar-

ray RF systems will of necessity involve some spiral evolution due to the rapid evolution (see Section 6) expected in the electronics capabilities. However, the existing model developed for the AMRFC testbed provides an excellent starting point. An example of the type of flexible modularity built into that system is the use of multiple channel receive and transmit modules. Design of a standardized interface design that will allow the number and type of such modules to be rapidly changed is an example of the type of standardization that will be essential to optimize development of multi-function phased array RF systems.

In developing hardware standards, the commercial standardization processes tend to be more formal than those described above for software standards. For example, the Institute of Electrical and Electronics Engineers (IEEE) is one of the leading standards organizations in the world, and through the IEEE Standards Association (IEEE-SA) has a process that has developed and promoted many of the standards in use today.

The IEEE process is more formal than the IETF process, and consists of several steps. The success of this process is unquestionable, as evidenced by the interoperability of hardware components used in modern electronics.

The IEEE standards process can be divided into seven steps:

1. A proposed standard must secure sponsorship from an IEEE-approved organization. This organization will be responsible for shepherding the proposed standard through the process. Examples of these IEEE-approved organizations are the various societies of the IEEE and their technical committees.

   In the case of the AWA system, the sponsor would be the entity that wants to promulgate a given standard.

2. In order to initiate a standards project, a Project Authorization Request (PAR) is submitted to the IEEE-SA Standards Board and it is reviews by the New Standards Committee which makes recommendations to the Board regarding approval of the PAR.

   In the case of the AWA system, this could be a steering committee put together by the Navy to manage the standards process. It would include representatives from the Navy as well as contractors.

3. A working group is then assembled. The rules of the IEEE-SA ensure that the working groups are open and that all qualified individuals have the right to attend and contribute to the meetings of the working group.

4. The working group prepares a draft of the proposed standard that follows the IEEE Standards Style manual.

5. The draft standard prepared by the working group is submitted to the IEEE Standards Department which sends an invitation-to-ballot to any individual who has expressed interest in the standard.

   In the case of the AWA system, this could be the same steering committee that has been put into place to guide the standards process.

6. If the balloting results in 75% approval then the draft standard is submitted to the IEEE-SA Standards Board Review Committee. This committee reviews the standard for compliance with IEEE Standards Board rules given in the IEEE-SA Standards Board Operations Manual, and recommends whether the standard should be approved.

7. The IEEE-SA Standards Board conducts a final vote on the proposed standard. A simple majority is required to approve the standard.

6 DESIGN PATHS

For multifunction apertures, the key design objectives are functionality, robustness and cost, where cost includes not only the initial purchase cost, but also the continuing operations and maintenance costs, and the costs of future upgrades. At present Multifunction Radio Frequency Apertures (MFRFAs) are planned to serve communications, radar and electronic warfare functions over a broad range of frequencies. These functions cover nearly all the RF needs of naval ships and we will focus discussion on the 2 to 20 GHz range. HF, VHF and UHF frequencies (3 MHz to 1 GHz) can be served by separate apertures of other antenna types, e.g., monopoles, helix and dish antennas.

In the design process, definition of the objectives and then requirements for a system are the key elements in the design process as diagrammed below in Figure 26. All the rest of the design flows from the objectives (broad goals of system) and the requirements for a specific application. A second key part of the design process is the definition of measures of effectiveness, such as cost, robustness against failures and damage, as well as functionality and timeliness.

Such a design process presumably guided the design for the AMRFC test bed. The multifunction apertures for Integrated Topside should be guided by the same design process. Engaging in a spiral design process will lead to the most effective selection of research activities most likely to have a high payoff in moving multifunction apertures into the Navy's fleet. The team that designed the AMRFC test bed has the skills to do this. As a step in the multifunction aperture design process for integrated topsides for the

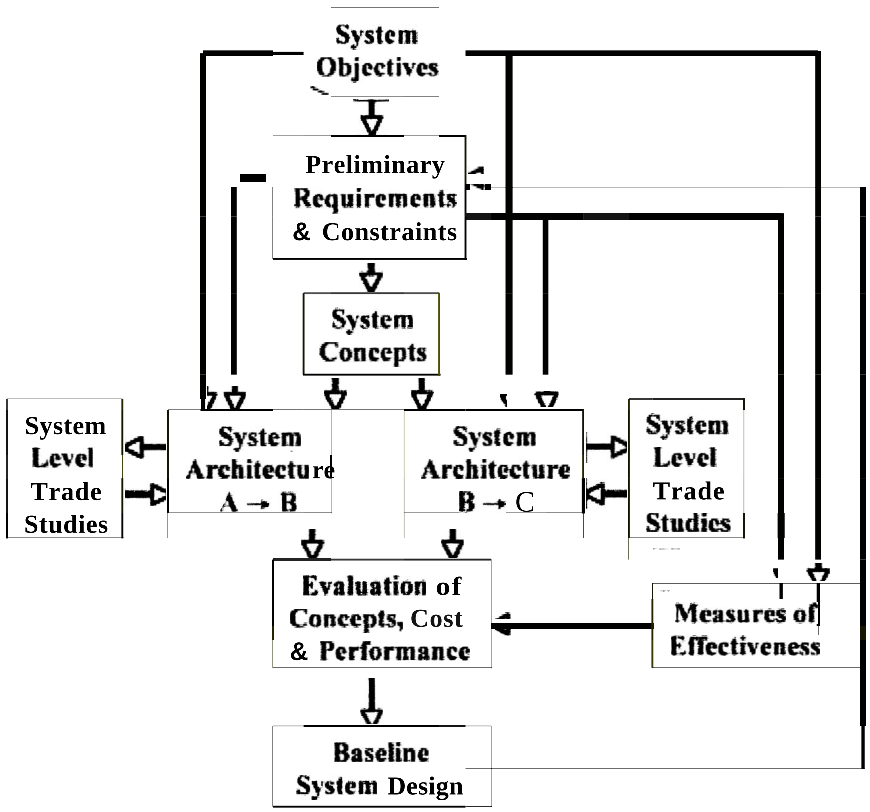


Figure 26: Block diagram of a system design process (adapted from Wertz and Larson [76]). Note how the result (baseline system design) of the design process feeds back to the requirements and constrains, as might happen if the baseline design satisfied the requirements, but not the cost constraint. The notations A $\rightarrow$ B (B $\rightarrow$ C) indicate that the feedback will be designed specifically to evaluate and integrate new technical capabilities in a spiral evolution.

Navy, we have identified three distinct design pathways (analogous to the alternative system architectures A,B, & C in Figure 26). We describe these paths below, characterizing them briefly and discussing their advantages and disadvantages.

The three design pathways that we considered are constructed from interaction with many sources of information and advice. Three distinct options have emerged as follows:

- Path A is a baseline path and features implementing currently available cost-saving technology into an AMRFC-like architecture
- Path B features "going with the flow" of mainstream technological advance dictated largely by adapting emerging commercial technologies
- Path C features advanced technology that requires developing and applying custom technologies specific to multifunction apertures.

In all these pathways the transmit and receive apertures are separated to achieve as much isolation as possible, with an achieved level of 80 to 100 dB.

To contrast the different pathways we will focus on the following features of the design:

- Location in the signal paths at which the transition between analog and digital signal flow occurs
- Method of beam forming and steering
- Method of accommodating the need for wideband linear RF amplifiers
- Likely cost considerations based on dual use to secure lower cost
- Performance and technological risk

Path A, currently available technology with AMRFC-like architecture: This path is the lowest risk and would produce a working system most quickly. This path would employ technology similar to the existing AMRFC test bed, modified to exploit low-cost manufacturing and COTS

components. A high level view of the AMRFC system architecture A is summarized in the following block diagram, Figure 27. **In** this design path the divide between digital and analog signals takes place as follows for the transmit and receive functions:

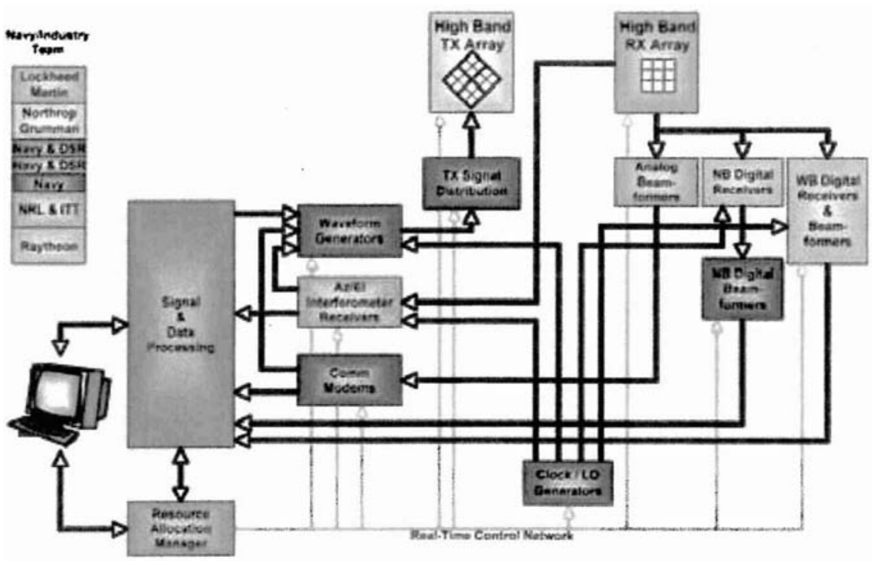


Figure 27: Block diagram of AMRFC test bed showing high-level system architecture. This characterizes design path A. (After Tavik et al., [4]).

- **Transmit:** Both wideband and narrow band signals are generated digitally (direct digital synthesis, DDS) at baseband (information bandwidth), converted to analog (DAC) and then multiplied and/or heterodyned up to the RF transmit frequency. Up to four different signals can be generated and distributed to any of the four quadrants of the antenna array. Each quadrant is assigned only one signal at one time, although time-multiplexing of two signals is allowed for EA application. Wideband waveforms up to 1 GHz bandwidth can be accommodated. Narrowband signals of $\approx$ 25 to 100 MHz bandwidth, but high spectral purity, are also generated by DDS, but in a separate path. The analog signals thus produced are used to modulate an optical signal and fiber optics are used to relay the analog signals to the

transmit modules in the array for amplification and transmission. Near the aperture the light signals are detected (producing an analog RF signal). The phase shifting to feed individual elements and accomplish beamforming is done after the conversion from light signals back to analog RF. The final RF amplifiers are class AB linear amplifiers with an efficiency of about 35-40%. This means that for high power applications, cooling must be provided to remove heat dissipation equal to the power transmitted - in addition to heat generated by other electronics near the aperture face. The highest power transmit signal is typically the air surveillance and tracking radar and this requires typically $\approx$ 5 to 10 kW average power. Thus, about 5 to 10 kW of heat would need to be removed over the area of the aperture due to losses in the RF amplifiers. The other transmit signals (see Table 1, Section 2) are much lower power, and cooling demands would be less severe for apertures dedicated to combinations of these lower-power functions.

• Receive: The receive array is divided into 9 sub-apertures containing 128 antenna elements each. The signal from each antenna is partitioned into four separate receive modules, allowing four fully independent beams or functions to be handled simultaneously. Beam forming is performed at the sub-aperture for each of the four functions [4]. The summed signals are then power-split, to allow processing in one of the three different paths shown in Figure 27. In the analog path, three, six or nine of the sub-aperture signals for each function are combined using wide-band power combiners. Demonstrations of communications and radar functions were performed using the analog path [4]. In the narrow-band digital path, a subset of the sub-aperture signals are downconverted to a 75 MHz IF and digitized to 14 bits. The digital signals are transmitted to the DSP via fiber optics for final beam forming. In the wide-band digital path, a sub-set of the sub-aperture signals

are downconverted to a 720 MHz IF (230 MHz bandwidth), and digitized at 8 bits. The second stage of beam forming, this time with with true-time delay, is performed at IF and the final beams are transmitted over fiber optic to the DSP.

The technology used in path A was designed to test functional performance and hence is relatively expensive. If more multifunction apertures were built using the path A approach one could expect some savings on design and on larger production lots for the array electronics. In addition, specific development of low-cost manufacturing processes, and the use of COTS components wherever possible would be needed to significantly lower cost. In overall system design (see Section 5), careful trade-offs involving limiting function on certain aperture elements, which would lower per-element costs, could be used to further limit costs.

We also note that a number of commercially driven materials developments offering improved performance for wider band applications and/or lower cost are presently in the engineering design and test stage, e.g., MEMS phase shifters, BST phase shifters and GaN amplifiers (see Section 3). Depending on the viability of these developments the Navy could leverage the commercial research investment in these devices for multifunction aperture considerations within the architecture of design path A.

However, the system design does use circuit architectures predicated on the design choices of GaAs MMICs. Achieving the very large cost reductions that may occur as RF-CMOS designs move into more demanding commercial applications would require substantial changes in the AMRFC architecture. The functionality of a path A design is impressive, but would need to be enhanced significantly to accommodate the cost and functional needs of an integrated topside, such as the Littoral Combat Ship (LCS).

The advantages of path A focus on the following:

- Demonstrated design, so technological risk is low.
- Limited research costs
- Early availability for use in fleet
- Potential for lowered costs with existing COTS, improved manufacturmg

The disadvantages of path A can be summarized as follows:

- Limited functionality relative to more advanced technologies of paths Band C
- Multifunction limited to switching between different sub-arrays
- Limited potential for advanced digital processing
- Final transmit signal amplification would be done by class AB linear amplifiers and suffer the 50% efficiency of this method

In summary path A is a near term, low risk option. Most of the research has been done and improvements over the AMRFC test bed would be in terms of lowering cost and enhancing functionality (more beams, more channels, etc.) by expansion based on demonstrated COTS technology. However, in terms of the paths A, Band C discussed here, path A would have the least functionality - though it would still represent a very significant step forward.

Path B features "going with the flow" of mainstream technological advance dictated largely by commercial applications: Commercial developments in **RF**-CMOS are changing the design rules that govern

performance and cost for the front end electronics in phased arrays. Path B seeks to take maximum advantage of these trends as in the notional design suggested in Section 4.1 and the commercial developments discussed in Section 4.3. Principal design features are illustrated in Figures 10 and 20 of Section 4 and summarized as follows:

- The cost advantages of CMOS technology fabrication, e.g., mixers, allow parallel processing of virtually all beams individually in parallel structures as illustrated in Figure 10.
- Parallelism allows use of lower resolution DAC's and ADC's
- Beamforming would be done primarily in the digital domain (after down conversion on receive and before up conversion on transmit)

On transmit the waveforms would be generated digitally and converted to analog at the intermediate frequency which would then be up converted to the desired transmit frequency and amplified for final delivery to the array antenna elements. Each function may need to be treated individually by use of parallel transmit modules. The phase-shifting (or true-time delay) would be done either at up-conversion from IF, or digitally before DAC, up conversion and final amplification.

On receive each beam or functional component of each antenna element would be treated individually with down conversion in parallel channels (viz. Figure 10 in Section 4). The beam forming would be done either by a digitally controlled VCO in the down conversion mixer or digitally, with true time delay, after the ADC. By separating the signal into functional channels individually at the antenna elements, the required resolution of the individual ADC's is relaxed and wider bandwidth signals can be accommodated.

The philosophy of path B is to use duplication of subsystems that are cheap in CMOS. Beamforming would be done at the intermediate frequency level either by using CMOS voltage controlled oscillators or in the digital domain. This approach allows one the benefits of the cost savings associated with CMOS technology fabrication.

The advantages of path B focus on the following:

- Lowering cost by taking advantage of the economics of CMOS fabrication and other processes rendered less expensive because of large-scale commercial markets
- Leveraging research investments aimed at the commercial market by using these parts for multifunction aperture applications
- Redundancy of many parallel channels allows easy remediation of failures in a single channel.
- Increased ability to use digital techniques e.g., for non-linear distortion in front-end processing of the waveforms.
- Relatively early introduction to the fleet

The disadvantages of path B can be summarized as follows:

- Possible increase in relative complexity due to the need to control many parallel channels
- Final transmit signal amplification would be done by class E tuned RF amplifiers which would limit bandwidth, without returning or multiple channes to the hundreds of MHz range

In summary path B is a medium term, moderate risk option. The notional design ideas of Section 4 above would be exploited to maximize the use of **RF** CMOS technology. This would allow cost reduction by the use of commercial fabrication facilities that can amortize costs over a large commercial market and must do so to be competitive. **In** terms of the paths A, Band C discussed here, path B would likely have more functionality than path A and less than path C, but the design process of Figure 26 would need to be exercised to determine the final outcome.

Path C features advanced technology that requires research specific to multifunction apertures: This path features advanced technology that is currently in the research and/or development stage and has decided advantages for multifunction **RF** apertures. The vision for Path C is to achieve a design that permits maximum functionality with the smallest amount of front-end electronics. This goal would be achieved by putting the digital to analog interface as close to the antenna elements as possible, and thus permitting the needed flexibility to control all the needed **RF** functions digitally. Further, the goal is to implement this plan with maximum efficiency to reduce the heat removal load on transmit. To accomplish this high level of performance a number of technological innovations and advancements are needed, for example:

- Specific design of ADC, DAC, beam summation and amplification electronics to match the requirements of active electronically steered arrays.
- Very high speed ADC and DAC functions to match the highest **RF** frequencies to be used by the aperture.
- High resolution ADCs to provide the high dynamic range needed for extremely wide bandwidths

- High efficiency, very linear, wideband power amplifiers for transmit, or an equivalent digital method based on delta-sigma DAC ideas.

Some of the technologies to implement this advanced vision would be extremely high-speed semiconductor electronics and designs and devices for direct digital synthesis at high output power. See for instance dicussions in Section 4.2.1 and Appendix D.

The advantage of this path is that it would, in principle, allow an almost completely digital RF system that could very flexibly employ the physical aperture to do all the needed functions without the complication of segmenting the system into frequency bands with the associated problems of scheduling and construction complexity. Such a multifunction aperture would allow the user to change whole methods of communications, surveillance and electronic warfare in software. This would enable Navy ships to adapt to the tactics of adversaries very rapidly and thus 'get inside their innovation loop.

The disadvantages of the approach of path C are the high technology risk, cost and timeliness. Pushing signal handling methods now in use in audio, ultrasound and digital imaging applications to the frequencies needed for active electronically steered arrays is inherently difficult and risky. Even if this succeeds, the cost of systems and devices that are specific to multifunction apertures will almost surely exceed the cost of path B in which commercial developments are leveraged. The time for this approach to find application in the fleet will be longer to accommodate the needed research and development. It would be unwise to focus solely on this approach. However it is essential to insure that the design path is flexible and that standardized interfaces and software architectures are used, so that future developments

in this high-risk/high-pay-off arena can be readily implemented into evolving Integrated Topsides systems.

Our recommendation: On balance and given the potential cost benefits of CMOS-RF front ends, path B with parallel transmit/receive processing elements for different RF functions at different frequencies is the most likely path to Naval implementation of wideband multifunction apertures. However, path C offers the prospect of the discovery and reduction to practice of advanced methods that could make multifunction apertures more effective and efficient, but probably with higher cost. The paths A, Band C should be thought of as a spiral development, in which the investments in each sequential stage are determined based on the results of the previous stage.

7 CONCLUSIONS

The development of flexible and adaptable active phased array systems, as envisioned in the Integrated Topsides Program, is essential to maintaining Naval superiority in radio frequency functions such as radar, communications and electronic warfare. **In** addition, the program is needed to control costs and address issues of radar cross section, fratricidal electromagnetic interference, and topside weight. The Integrated Topsides concept has outstanding potential to address all of these issues, because technology developments directly applicable to program needs are now available, with rapidly improving capabilities.

A key element in developing the next generation of active phased array technology rests in high-speed, high-power semiconductor electronics. The Department of Defense has played a key role in supporting the development of fast and high-power semiconductor materials and devices, which in turn have played a major role in the development of commercial technologies such as wireless communications. As a result, there is now a dynamic commercial enterprise involved in rapid development of new and relevant **RF** capabilities. This gives the DoD the opportunity to recoup some of its initial investment by exploiting the efficiencies of scale in commercial components. **In** addition, the new technologies will allow a rapid transition from the historical analog signal processing used in phased array systems, to more powerful digital approaches. this turn provides additional cost benefits by allowing software-based detection, calibration and correction of hardware weakness. We have addressed specific opportunities in relevant technologies in Sections 3 and 4, and detailed technical findings are presented at the end of each of those sections.

A second key element in developing the multi-functional arrays of the Integrated Topsides Program is management of the complexity of the overall system. The Navy has supported an excellent technology demonstration in the AMFRC program, which has demonstrated the principles of the necessary management. These include the care in design and placement of multi-function array configurations to avoid interference, and to allow rapid test and calibration. Also included is the use of open architectures and standardized interfaces to allow ready reconfiguration, adaptation and integration of new technological capabilities. This adaptability is also important in insuring robust system operation with respect to failure of individual components, as well as adaptability to changing battlefield environments. In addition, the AMFRC program demonstrated dynamic priority-based allocation of the resources of the multi-functional arrays to the different functions depending on the tactical environment. Continuing with and formalizing these practices, as described in Section 5, under the direction of cognizant Naval personnel is required for fully realizing the potential of the Integrated Topsides Program.

The importance of open, adaptable hardware and software interfaces cannot be overemphasized: Because the technology base is evolving rapidly, immutable decisions about the system configuration should not be made early in the program. Instead, a spiral development program should be implemented, as described in Section 6. The first coil of the spiral would use immediately available technology to develop low-cost prototypes that demonstrate one or more aspects of the entire desired functionality. Based on the success and lessons of the prototypes, more advanced commercial technology developments would be used to guide design of systems with improved multiple-function and well-controlled cost. Increased use of digital processing will play a key role in improving function while limiting cost. Finally, the

results of long-term applied research and development on highly specialized capabilities should be incorporated in the system subject to the cost and benefits of improved function.

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A APPENDIX: Briefings

Tour of the Advanced Multifunction RF Test-Bed	Betsy DeLong, ONR and Greg Tavik, NRL
Multi-function Arrays and Integrated Topside	Bobby Junker, ONR
Agile Wide Band Arrays	Peter Moosbrugger and Keith Kelly Ball Aerospace
Architectures and Component for Advanced Multifunction Arrays	Leo PelIon and **Wil** Lew Lockheed-Martin
Multifunction RF System Overview	Michael Fitelson, Northrop Grumman
SATCOM and Line of Sight Architecture	John Przybysz, Northrop Grumman
MFEW Program Overview	Marvin Hunter, Northrop Grumman
Heterogeneous Chip	Nicholas G. Paraskevopoulos, Northrop Grumman
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Raytheon RF Multifunction Technologies	Tom Markarian, Raytheon
Affordability and Scalability/ Modularity of Active Apertures	Mike Sarcione, Raytheon
Open System Architure for Surface Radar	Robert Kingan, Raytheon
Wide Bandgap (GaN) Technology at Raytheon	Joe Smolko, Raytheon
Si RF CMOS SoC Technology for Low-Cost AMRFC	Dick Healy, Raytheon
Digital Receiver and Exciter (DREX)	Shamsur Mazumdar, Raytheon
Overview of the Advanced Multifunction RF Concept Test-Bed	Greg Tavik, NRL
Wideband Wavelength-Scaled Antenna Concept	Mark Kragalott, NRL
Mixed-Signal Technologies for microwave phased arrays	Mark Rodwell, UCSB
Wideband Agile Arrays Technologies Concepts and Recommendations	Steve Hedges, BAE Systems

B APPENDIX: Acronyms

AAW	anti-air warfare
ADC, DAC	analog-to-digital, digital-to-analog converter
(A)ESA	(active) electronically steered array
AMRFC	advanced multi-function RF concept
BER	bit error ratio
BFN	beam forming network
BJT	bipolar junction transistor
CDL, TCDL	common data link, tactical CDL
COTS	commercial off-the-shelf
DDS	direct digital synthesis
DFT	Discrete Fourier Transform
DSP	Digital signal processor
EIRP	effective isotropic radiated power
EW, ES, EA	electronic warfare, support, attack
FIR	finite impulse response
FSS	frequency selective surface
GaN,BST	gallium nitride, barium strontium titanate
HBT	heterojunction bipolar transistor
HFET	heterojunction FET
HGHS	high-gain high-sensitivity
HPOI	high probability of intercept
IF	intermediate frequency
ILS	instrument landing system
IMD, IMP	intermodulation distortion, 1M product
LCS	Littoral Combat Ship
LNA, PA	low noise amplifier, power amplifier
La, I, Q	local oscillator, in-phase, quadrature
LPF,BPF,HPF	low, band, high-pass filter
MAC	media access controller
MEMS	micro electromechanical systems
MIMO	multiple input/ multiple output
MMIC	monolithic microwave integrated circuit
MUX, DEMUX	multiplexer, demultiplexer
OACE	open architecture computing environment
OFDM	orthogonal frequency division multiplexing
PDF	precision direction finding
Pll	phase-locked loop
RCS	radar cross section
ROSA	radar open system architecture
SFDR	spurious free dynamic range
SIP	system in a package
SOC	system on a chip
T/R (TRM)	transmit/receive (module)
Tx, Rx	transmitter, receiver
UWB	Ultra Wide Band
VSWR	voltage standing wave ratio
WCDMA	wideband CDMA
WiFi	wireless fidelity (IEEE 802.11 network)
XPDR	transponder

C APPENDIX: New Approaches for Waveform Generation

C.l Using Chaotic Waveforms and Nonlinear Synchronization for High Bandwidth Radar Systems

There is a long history of exploring waveforms in radar systems which are "noise." **In** such a system the antenna in a radar transmitter produces a waveform $s(t)$ which is selected from a time series of a noise signal. Noise is defined to be a signal which has autocorrelation unity at $t = 0$ and autocorrelation zero at other times. An example of this is given in Dawood and Narayanan [77], and the idea goes back at least to Cooper and Gassner [78].

The receiver can extract the return signal from this transmitted waveform by correlating the original waveform ("noise") with the return, and using standard methods make a decision whether the received signal is, in fact, the return from a target.

The advantages of such radar waveforms is that the bandwidth is large so that time resolution, and thus range resolution, should be very good, and the ambiguity function (Woodward [79]) can, in principle, be very sharp. The latter comes from the fact that the area under the ambiguity function is unity, and the sidelobes are of order $\frac{1}{TB}$, with T the transmit time and B the bandwidth in the transmission. This can be quite small. See figure from the presentation by Harman [81] at the workshop http://www.enm.bris.ac.uk/workshop/chaotic-signals/.

In the past decade, there has been a new twist on this idea beginning

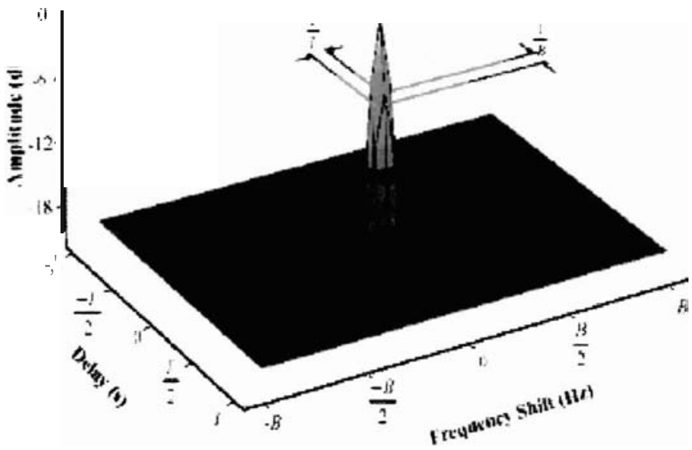

Figure 28: Thumbtack ambiguity function, as achieved by a noise waveform.

with the suggestion of Bauer [80] to substitute for the "noise" waveform discussed earlier, a wave form generated by a chaotic dynamical system. Such a waveform is also wideband, giving the advantages just discussed, but it also is easily generated using low power simple devices. The chaotic signal can be generated in analog hardware or in a very simple fashion on a DSP or on a FPGA as simple sets of ordinary differential equations will do the job.

As pointed out by Harman [81] and others, another advantage of chaotic waveforms is that the transmitter need not be complex or made expensive by the need to have elaborate feedback circuits for linearizing the signals.

One can transmitter more than one waveform if orthogonal samples in the transmit interval T can be found. To establish a set of signals orthogonal over the time interval T of a transmission, on can utlilize function orthogonal on the attractor of the nonlinear system producing the chaotic signal. For this purpose consider the state of the chaotic system generating the signal to be transmitted $\frac{1}{\tau}$ as described by the D-dimensional vectors $y(l)$ sampled at frequency. These vectors are $y(l) = y(t_o + (l - 1)7); l = 1, 2, \ldots, N; NT = T$

with T the time over which the chaotic system is sampled. The natural measure on the attractor is

$$p(x) = \frac{1}{N} \sum_{l=1}^{N} \delta^D(x - y(l)).$$

And we seek functions $\varphi_m(x)$ such that

$$\int d^D x\, p(x)\varphi_m(x)\varphi_k(x) = \delta_{mk}$$

$$= \frac{1}{N} \sum_{l=1}^{N} \varphi_m(y(l)\varphi_k(y(l))$$

These can be constructed in a "Gram-Schmidt" fashion, starting with $\varphi_0(x) =$ 1 and continuing on using polynomials or other functions in x space. Once we know the $\varphi_m(x)$, we can generate $y(l)$ from the chaotic system and transmit one of the $\varphi_0(x) =$ 1 over the allotted time interval T.

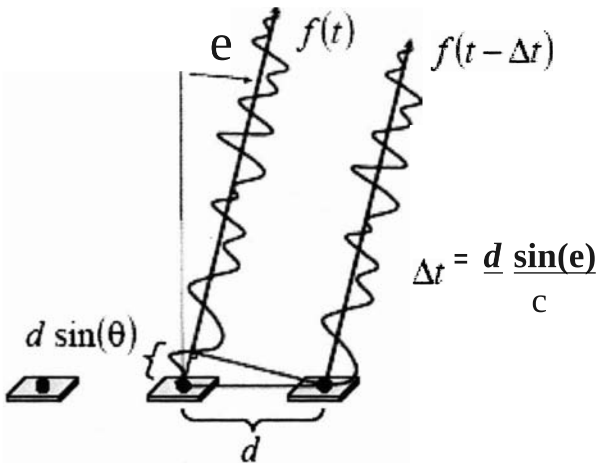

Figure 29: Beam steering in a wide bandwidth array requires a true time delay.

Another issue in chaotic radar is the construction of arrays where the signal transmitted is delayed appropriately from one element to another so that a good beam is formed. This is illustrated by the figure above.

To create the appropriate time delay between elements, Corron, Blakely, and Pethe [82] have suggested using the phenomenon of **lag synchronization** discussed by Rosenblum, Pikovsky, and Kurths, 1997. This is a phenomenon where slightly mismatched chaotic oscillators synchronize with $f(t)$ the signal in one and $f(t - \Delta t)$ the signal in the other. By creating a chain of such lag synchronizing systems, and inducing the lag by the adjustment of a capacitor in an electrical circuit realizing the chaotic oscillator, they are able to show (ibid, 2005) that beams can be formed rather easily and steered by a master control of the appropriate capacitors in the array elements.

These ideas represent novel directions which may be of interest to the sponsor. Some of the work is done at the Redstone Arsenal of the Army, so it should be readily available for Navy applications.

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